WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

AUGUST 1951

VOL. XXVIII No. 335 THREE SHILLINGS AND SIXPENCE

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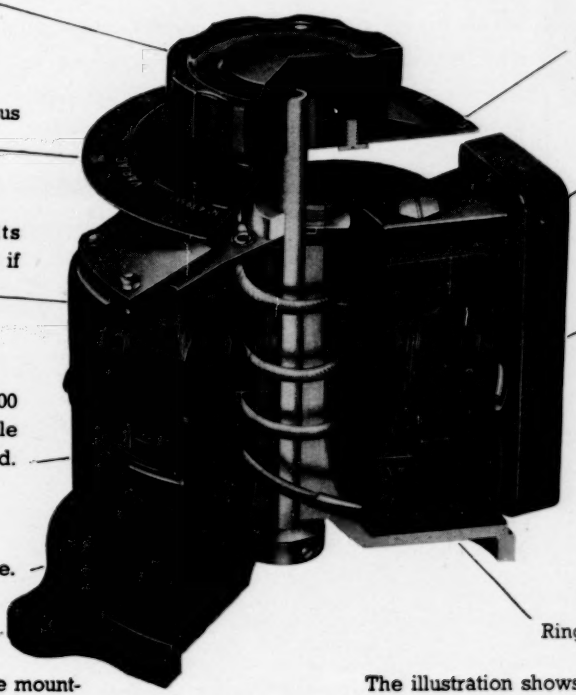

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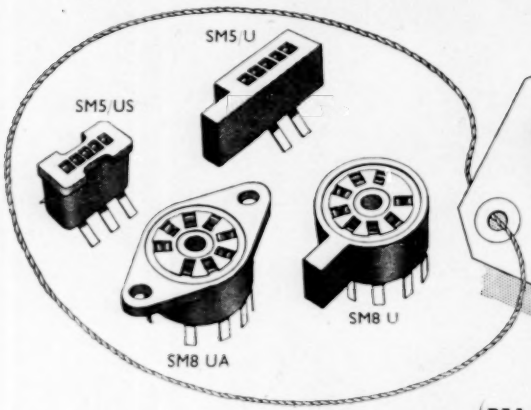

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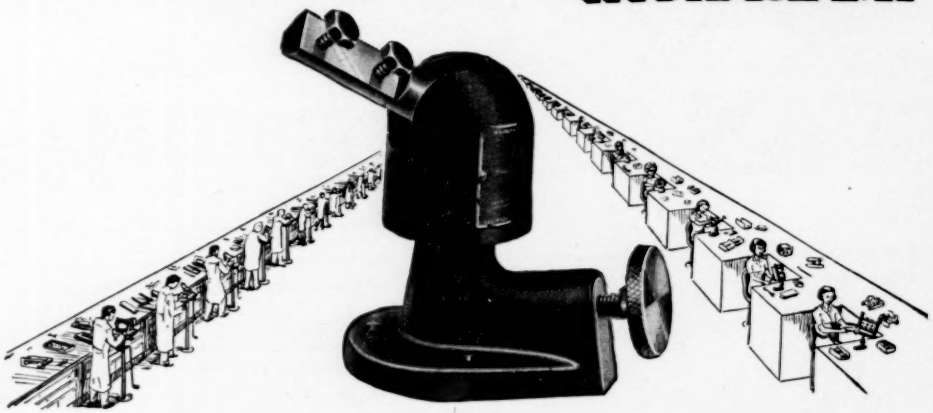

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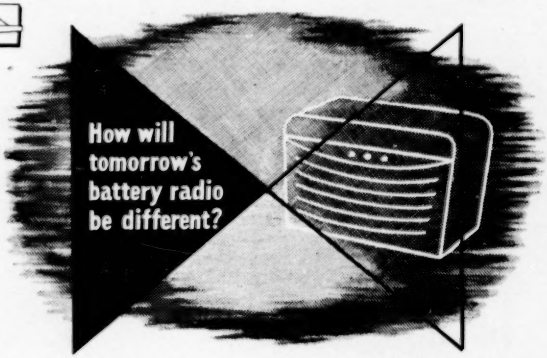

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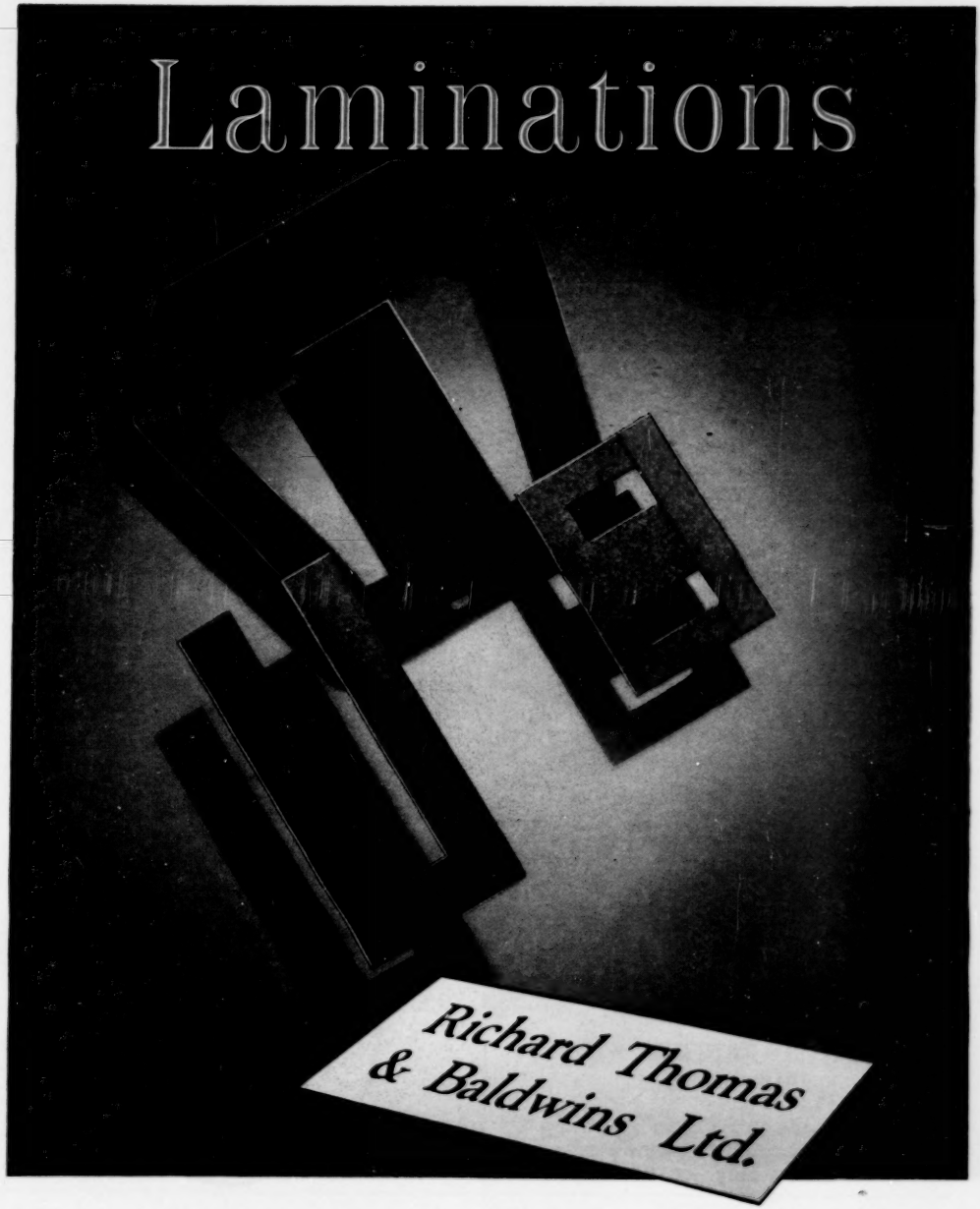

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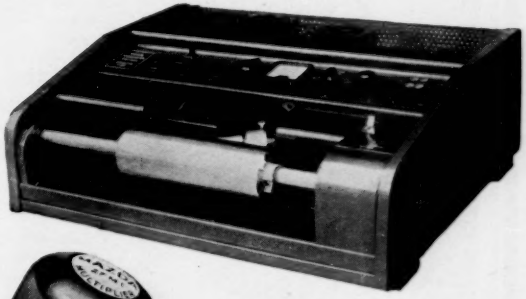

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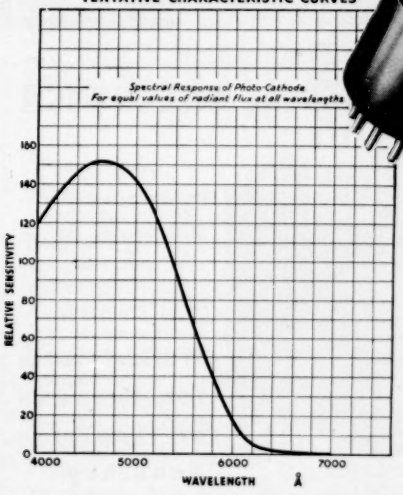

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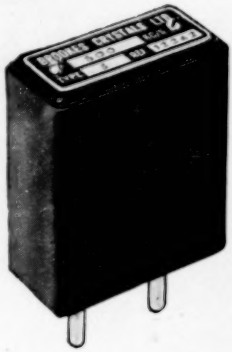

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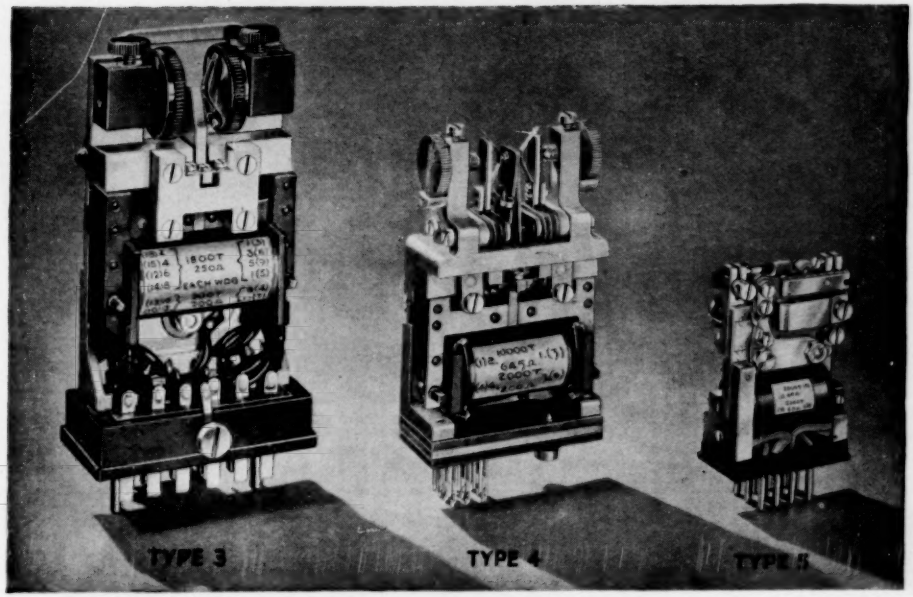

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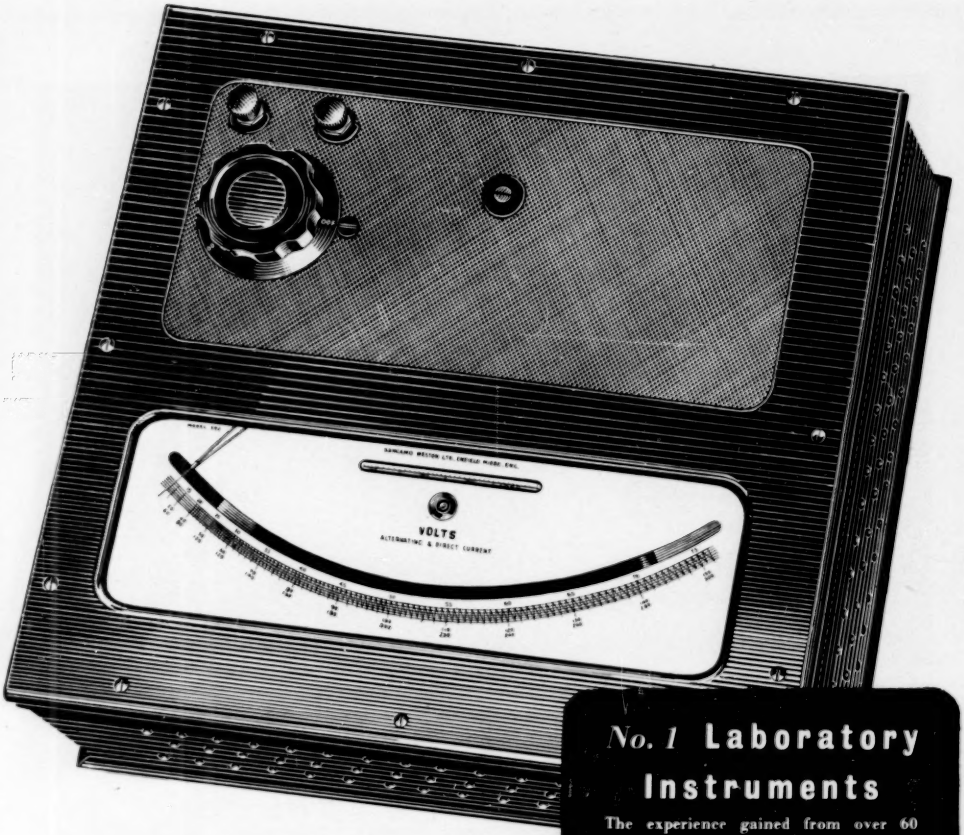

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AUGUST 1951

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VALVE DATA	
Nominal Operating Voltage	85V
Max. Starting Voltage	125V
Current Range	1-10mA
Preferred Operating Current	6mA
Internal Resistance at 6mA	300 Ω

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An LR Load Circuit

SUPPOSE that a resistance R and an inductance L are connected in parallel and supplied with current by a generator of infinite internal resistance. Let the instantaneous currents in R and L be i and i_s and the total current be i_T, while the peak-to-peak current in R is I. One is inclined to say that the total current must always be greater than the current in R by the amount existing in L. This is not always the case, however, for the inductive element can store energy over one part of a cycle and restore it over another; as a result, one can have a condition in which a comparatively low value of L produces greater circuit efficiency than a high one.

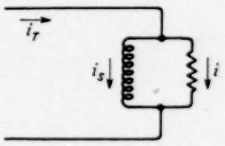

Fig. 1.

As a case in point, suppose that it is required to produce in R a repetitive saw-tooth waveform with a linear scan stroke of period τ and of negligible fly-back time. The current in R over any one cycle when the steady state has been reached can be written as

$$i = I(x - \tfrac{1}{2}) \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where $x = t/\tau$.

The voltage across R is $e = iR$ and the current in L is $i_s = \frac{\tau}{L}\int e\,dx = \frac{\tau R}{L}\int i\,dx = 2y\int i\,dx$ and, since $\int_0^1 i_s dx = 0$,

$$i_s = Iy[(x^2 - \tfrac{1}{3}) - (x - \tfrac{1}{2})] \quad .. \quad .. \quad (2)$$

whence

$$i_T = I[y(x^2 - \tfrac{1}{3}) + (1 - y)(x - \tfrac{1}{2})] \quad .. \quad (3)$$

This expression has a minimum value when $x = \frac{1}{2} - 1/2y$, but this only applies for $y > 1$ since x must be positive. When $y \leqslant 1$ the minimum occurs for $x = 0$.

Let us consider this latter case first. When $x = 0$, $i_T = i_{T(0)} = -\frac{1}{2}I(1 - y/3)$. The peak-to-peak current is the current at $x = 1$ less the current at $x = 0$ and is I; it is quite independent of y.

Now if the generator is a valve, it is necessary that its current shall never be negative. It must, therefore, have a direct-current component of minimum magnitude $-i_{T(0)}$ which will, in fact, be the minimum mean anode current i_a of the valve. Therefore,

$$i_a = \tfrac{1}{2}I(1 - y/3) \quad .. \quad .. \quad .. \quad (4)$$

and the peak current of the valve is clearly

$$i_p = I \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

As y is varied from zero to unity the peak current remains unchanged but the mean anode current varies from $I/2$ to $I/3$. Now $y = 0$ means $L = \infty$, which is usually considered to be an ideal condition, but an impracticable one. However, $y = 1$, or $L = \tau R/2$, gives a finite value of inductance which requires only two-thirds of the mean anode current. Under this condition equation (3) reduces to

$$i_T = I(x^2 - 1/3) \quad .. \quad .. \quad .. \quad (6)$$

and the instantaneous anode current of the valve is x^2I. A linear saw-tooth can be obtained in R in spite of the presence of the inductance if the valve provides a square-law current; the mean anode current is reduced and the peak anode current is not increased. There is consequently considerable advantage to be gained by using a properly chosen value of L.

Physically, what happens is that when the fly-back occurs the current in L cannot change instantly. This current is $Iy/6$ [equation (2) for $x = 0$ or 1]. At the instant of fly-back it is being supplied by the valve; immediately after it is produced by the energy stored in the magnetic field and the current flows into R and so reduces the current which need be supplied to it by the valve. Towards the end of the scan cycle L robs R of current but over the first part it returns current to the circuit. The reason for the reduction of the mean anode current is simply a result of a change of waveform.

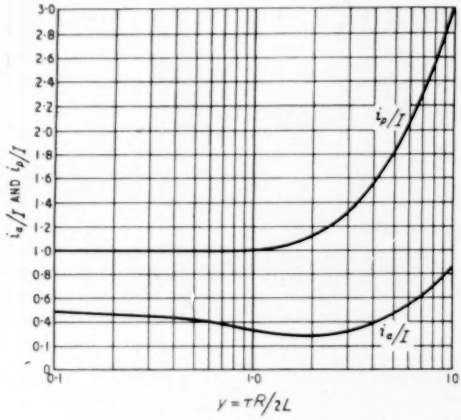

Fig. 2.

When $y > 1$ the current is no longer a minimum at the start of the cycle but occurs when $x = \frac{1}{2}(1 - 1/y)$. Arguing on the same lines as before, we find

$$i_a = \tfrac{1}{4}I(y/3 + 1/y) \quad .. \quad .. \quad .. \quad (7)$$

and

$$i_p = I(\tfrac{1}{2} + y/4 + 1/4y) \quad .. \quad .. \quad (8)$$

When y exceeds unity the peak anode current of the valve exceeds I so that it may not pay to increase y very much. The mean anode current has a minimum when $y = \sqrt{3}$ and it is obvious that y should not exceed this figure. With it $i_p = 1{\cdot}076I$ so that the increase of peak current is very small.

Values of i_a and i_p are plotted in Fig. 2 for a range of values of y and it is clear that $y = \sqrt{3}$ should normally be a maximum since the peak current increases rapidly for higher values. Generally speaking, for economy y should be between 1 and $\sqrt{3}$.

In practice, R might be 5000 Ω and $\tau = 0{\cdot}02$ sec whence $L = 0{\cdot}02 \times 5000/2 = 50$ H if $y = 1$. If an attempt were made to utilize a linear supply current, then to make the current in L depart from linearity by only a small amount, say 2%, it would be necessary to make L/R about 50τ or $L = 50 \times 0{\cdot}02 \times 5000 = 5000$ H. This is quite an impracticable value.

There is, of course, nothing new in the use of an inductance around 50 H only, when the 'ideal' requirements call for one of 5000 H. It is done in the frame time base of every television receiver. The facts that it leads to a reduction of mean anode current and a negligible increase of peak anode current are not too well-known, however, and are not easily determined by experiment because the so-called ideal conditions can hardly be realized. It has been previously referred to,[1] but in an article which dealt mainly with a method of producing the required grid-voltage wave for the valve. There are several ways of doing this; one common method is to use negative current feedback with the circuit discussed in the June Editorial.

The reduction of mean anode current consequent upon the use of a moderate value of inductance shunting a resistive load is hardly a result which one would expect and well illustrates the advantage of what may be called steady-state transient analysis, if this is not a contradiction in terms. By it we mean the analysis of the performance of the circuit to a repeated particular waveform for the condition that so many cycles have elapsed that successive cycles are identical. This form of analysis is often very difficult but there are certain cases in which, by a form of backwards working, valuable results can be obtained very easily. The circuit discussed here represents one such case.

W. T. C.

[1] "Projection Television Receiver," by J. Haantjes and F. Kerkhof, *Philips Technical Review*, April 1949, Vol. 10, p. 307.

CATHODE-FOLLOWER INPUT IMPEDANCE

Effect of Capacitive Load

By J. E. Flood B.Sc. (Eng.)

1. Introduction

THE advantages normally claimed for the cathode-follower are its high input impedance and its low output impedance: the latter enables it to maintain uniform amplification over a wide frequency band when working into a capacitive load. When the cathode load is capacitive, however, feedback through the cathode-grid capacitance can cause the input conductance of the stage to become negative at high frequencies.[1,2,3] This will cause oscillation if the cathode-follower is connected across a circuit which resonates at a frequency at which the negative input conductance of the cathode-follower has a greater magnitude than the positive conductance of the circuit across which it is connected.

The analysis and experimental results below determine how the input conductance varies with circuit parameters and with frequency. It is shown that the input conductance is modified by the Miller effect introduced by the unavoidable anode load. Various possible methods of avoiding negative input conductance are considered. The use of a series grid-stopping resistance is the only useful method for preserving a low positive input conductance at high frequencies; it has accordingly been analysed quantitatively.

2. Analysis of the Circuit

Fig. 1 shows an analysis of a simple cathode-follower circuit on the lines of Lockhart[4] and and Jeffery[5]. The equivalent input impedance of the circuit of Fig. 1(a) is shown in Fig. 1(c). When the cathode load is purely capacitive the equivalent input circuit contains a negative resistance and when the cathode load is purely inductive the equivalent circuit contains a positive resistance as shown at (f) and (g).

The input admittance of the circuit of Fig. 2 is deduced in Appendix 1. In this circuit the grid resistor is connected to a point in the cathode-load resistance (in order to provide a suitable bias voltage) and the cathode is capacitively loaded. It is assumed that the valve is working on the linear part of its characteristic and that the current in the grid resistor is small compared with the cathode current.

MS accepted by the Editor, August 1950

The input conductance (G_i) is

$$G_i = \frac{1}{R_g} \cdot \frac{(1 + g_mR_1)(1 + g_mR_k) - \omega^2R_kR_gC_{gk}(g_mR_kC_L - C_{gk})}{(1 + g_mR_k)^2 + \omega^2(C_{gk} + C_L)R_k^2}$$

and the input capacitance is

$$C_i = C_{ga} + C_{gk}\frac{(1 + g_mR_k) + \omega^2C_L(C_{gk} + C_L)R_k^2}{(1 + g_mR_k)^2 + \omega^2(C_{gk} + C_L)^2R_k^2}$$

where $R_k = \dfrac{(R_1 + R_2)r_a}{R_1 + R_2 + r_a}$

The input conductance is zero at frequency $\omega_o/2\pi$ where

$$\omega_o^2 = \frac{(1 + g_mR_1)(1 + g_mR_k)}{R_gR_kC_{gk}(g_mR_kC_L - C_{gk})}$$

At frequencies below $\omega_o/2\pi$ the input conductance is positive, but at greater frequencies it is negative. When $C_{gk} > g_mR_kC_L$, ω_o^2 is negative, therefore, ω_o is imaginary and the input conductance is positive at all real frequencies.[1] When $C_{gk} = g_mR_kC_L$, ω_o is infinite and it can be shown[6] that the gain of the circuit is independent of frequency and the phase shift is zero.

Since the anode of a cathode follower cannot usually be earthed except through a decoupling capacitor of finite susceptance, the input impedance of the circuit is modified by feedback through the anode-grid capacitance (the Miller effect). The component of the input conductance of a valve caused by a capacitive anode load is positive, so the frequency at which the input conductance of the cathode follower becomes negative is increased. Appendix 2 shows that if a cathode-follower has its anode connected to earth by a capacitance C_A, the input conductance is zero at frequency $\omega_o'/2\pi$ where

$$\omega_o' = \omega_o\sqrt{1 + \frac{g_mR_gC_{ga}}{(1 + g_mR_1)C_A}}$$

and $\omega_o/2\pi$ is the frequency of zero input conductance when the anode is earthed through zero impedance.

3. Methods of Avoiding Negative Input Conductance

3.1. *Effect of Circuit Values*

The input conductance cannot become negative if $C_{gk} > g_m R_k C_L$. Reducing g_m or R_k in order to satisfy this inequality reduces the gain of the circuit, reduces the maximum input voltage which can be handled without overloading, and reduces the input impedance of the circuit at low frequencies. Reducing C_L is usually impossible because it is already at the irreducible minimum consisting entirely of stray capacitances. Increasing C_{gk} is undesirable as it increases the low input capacitance of the cathode follower, which is normally one of the chief reasons for its use. If $C_{gk} > g_m R_k C_L$, then the input capacitance $> C_{ga} + C_L \dfrac{g_m R_k}{1 + g_m R_k}$ which may be greater than C_L.

3.2. *Anode-Grid Feedback*

Feedback through the anode-grid capacitance of a valve with a capacitive anode load produces a positive input conductance which may cancel the negative conductance due to feedback through the cathode-grid capacitance when the cathode load is capacitive.[1]

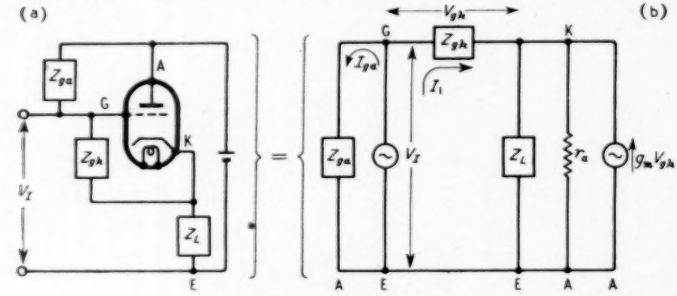

Fig. 1. *The circuit of the cathode follower is shown at (a) with its equivalent at (b). This can be further reduced to the form (c). When the cathode load is a pure resistance and $r_a \gg R_L$ the input impedance is equivalent to (d) and (e) at low and high frequencies respectively. When the cathode load is a pure capacitance and $r_a \gg 1/\omega C_L$ the input impedance has the form (f). When the cathode load is a pure inductance and $r_a \gg \omega L$, the input impedance has the form (g).*

$$V_I = I_1\left(Z_{gk} + \frac{r_a Z_L}{r_a + Z_L}\right) + g_m V_{gk}\frac{r_a Z_L}{r_a + Z_L}$$

BUT $V_{gk} = I_1 Z_{gk}$

$$\therefore\ Z_1 = \frac{V_I}{I_1} = Z_{gk} + g_m Z_{gk}\frac{r_a Z_L}{r_a + Z_L} + \frac{r_a Z_L}{r_a + Z_L}$$

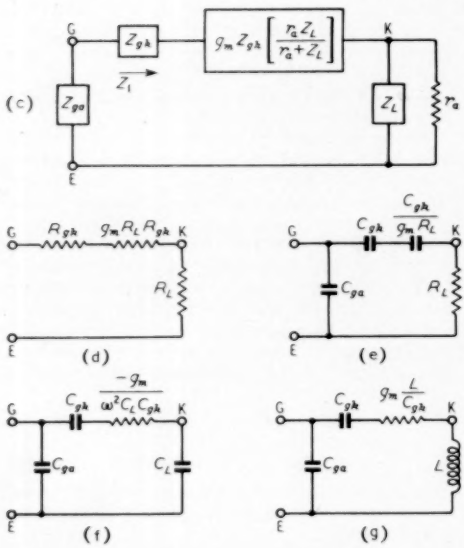

Fig. 2. *The circuit of Fig. 1(b) is redrawn as a bridge circuit in (b). When $C_{gk}C_A = C_{ga}C_L$ and $R_D C_A = R_L C_L$, the current which flows between points G and E, due to the generator μV_{gk}, is zero and the input impedance is that of the circuit of (c). The input conductance is, therefore, always positive, but may be large. If the input conductance is to be small at high frequencies, R_D and R_L must be either small or large compared with the reactances of C_D and C_L at the frequencies considered. If R_L is small, the gain of the circuit is small and if R_D and R_L are large, the h.t. supply voltage must be much larger than the anode voltage required by the valve.*

3.3. *Cathode Load Inductor*

Because a capacitive cathode load causes a negative input conductance and an inductive load causes a positive input conductance some advantage might be obtained by the use of a cathode load containing both C and L. But the capacitive load has the effect of connecting a resistance equal to $-g_m/\omega^2 C_L C_{gk}$ in series with C_{gk} [Fig. 1(f)] and the inductive load produces a resistance $g_m L/C_{gk}$ in series with C_{gk} [Fig. 1(g)] so the resistances can, therefore, cancel out only at one particular frequency, the resonant frequency of C_L and L. Above this frequency, the input conductance is always negative; below it

the input conductance is increased. Figs. 6 and 8 show that an inductor in series with the cathode resistor can compensate the high-frequency gain of the circuit and cause the input conductance to

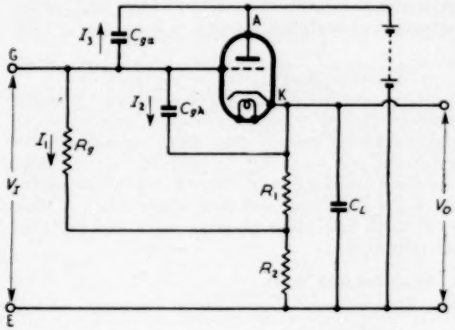

Fig. 2. *Cathode follower with grid leak returned to a tapping on the cathode load resistance.*

shunt resistance results in a low input resistance at low frequencies. The method is, therefore, ineffective.

3.5. *Series Grid Resistance*

At frequencies at which the input conductance is negative, the input impedance of the circuit with a capacitive cathode load is equivalent to a capacitance in series with a negative resistance, both being frequency dependent. A grid-stopping resistor can place an equal positive resistance in series with the negative resistance. The value of series resistance required is smaller at high frequencies than at frequencies within the bandwidth of the circuit, so a value which is large enough to prevent negative resistance over the required frequency band will prevent the input resistance becoming negative at any higher frequency. Because the grid stopper is a low resistance in series with the input impedance of the circuit, it will have little effect on the input resistance at low frequencies. The use of series resistance is,

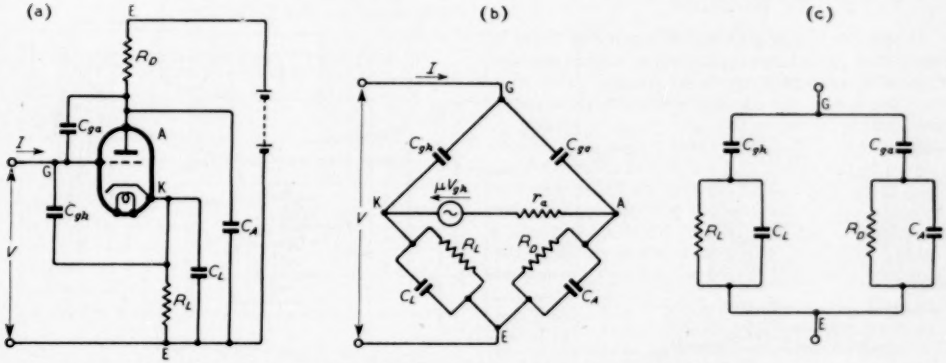

Fig. 3. *Cathode follower with anode-grid feedback (a). The circuit has the bridge equivalent (b) and when this is balanced its input impedance has the form (c).*

change sign at a slightly higher frequency, while increasing the conductance at lower frequencies.

3.4. *Shunt Input Resistance*

Figs. 4 and 5 show that, at frequencies at which the input conductance is negative, its magnitude increases with frequency. In order to avoid negative conductance over a given frequency band, at the upper end of which it occurs, the input must be shunted by a resistance as small as the arithmetic value of the input resistance at the highest frequency in the band. (A similar result can be obtained by reducing the value of the grid resistance). At frequencies above the band the input conductance is still negative and the added

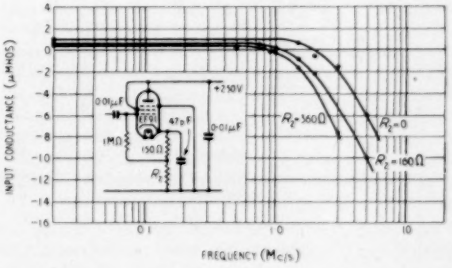

Fig. 4. *Effect of cathode resistance on input conductance.*

therefore, preferable to the use of shunt resistance.

It is shown in Appendix 3 that, if the frequency (ω_0) at which the input conductance is zero is small compared with the bandwidth of the stage, the required value of grid-stopping resistance (R_s) is approximately given by

$$R_s = 1/\omega_0^2 C_i^2 R_I$$

where C_i is the input capacitance of the circuit and R_I is the input resistance at low frequencies.

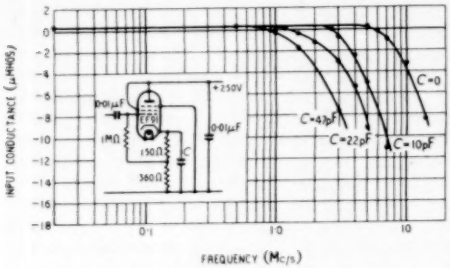

Fig. 5. *Effect of cathode-circuit capacitance on input conductance.*

When the screen grid (g_2) of a pentode valve is connected (by a large capacitance) to the cathode, C_{g1a} is small and C_{g1g2} is in parallel with C_{g1k} thus increasing the effective grid-cathode capacitance and reducing ω_0 (if $C_{gk} + C_{g1g2} < \frac{1}{2}g_m R_k C_L$); this increases R_s. When the screen grid is connected to the anode, C_{g1g2} is in parallel with C_{g1a}, thus increasing C_i and causing the Miller effect to increase ω_0; the increases in both C_i and ω_0 cause a reduction in R_s. When triode-connected, a valve, therefore, requires a much smaller grid-stopping resistance than when pentode-connected.

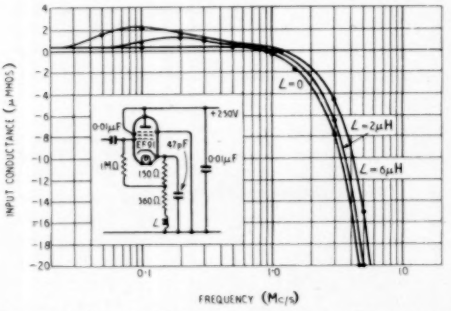

Fig. 6. *Effect of inductance in the cathode load on input conductance.*

A resistance in series with the valve causes the actual input voltage to the valve to decrease with increasing frequency; it will have decreased by approximately 3 db when $\omega C_i R_s = 1$. The gain of the circuit itself will have decreased by approximately 3 db when $\omega(C_L + C_{gk})\dfrac{R_k}{1 + g_m R_k} = 1$. Therefore, if the grid stopping resistance is not seriously to worsen the attenuation and phase distortion at high frequencies, it is necessary that $R_s < \dfrac{C_L}{C_i} \cdot \dfrac{R_k}{1 + g_m R_k}$. The value of grid-stopping resistance required to maintain the input conductance of a pentode valve positive at all frequencies is usually too large to satisfy this condition, whereas that for a triode is not. It is, therefore, preferable to use a triode-connected valve for a cathode-follower stage when a wide bandwidth and a low positive input conductance are required.

4. Experimental Work

4.1. *Measurements on Triode-connected EF91 Circuits*

Measurements of the input conductance-frequency characteristics of some cathode-follower circuits using a triode-connected EF91 valve were made by means of a Q-meter.

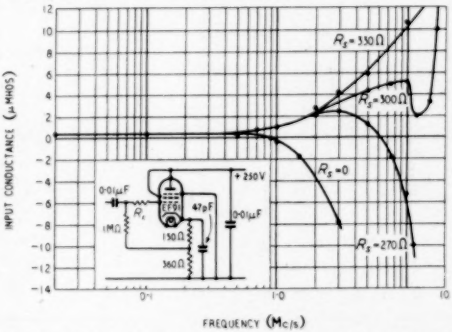

Fig. 7. *Effect of series grid stopper on input conductance.*

A coil was connected to the Q-meter and tuned to resonance at the chosen frequency ($\omega/2\pi$). The tuning capacitance (C) and the reading of Q were noted. If the dissipation of the coil is represented by a conductance (G_o) in parallel with a perfect inductance L, then $Q = 1/XG_o$ where $X = \omega L = 1/\omega C$

$$\therefore G_o = \omega C/Q$$

The input terminals of the circuit whose input conductance (G) was to be measured were connected across the coil and the tuning capacitor readjusted to resonate the circuit. The new magnification reading (Q') was noted. Then $Q' = 1/\{X(G + G_o)\}$

$$\therefore G = \frac{1}{X}\left(\frac{1}{Q'} - \frac{1}{Q}\right) = \omega C.\frac{Q - Q'}{QQ'}$$

If $Q' < Q$, G is positive. If $Q' > Q$, G is negative. If $Q' = Q$, G is zero.

The input conductance was measured for different values of cathode resistance, load capacitance, series cathode inductance and series grid resistance. Figs. 4 to 7 show the results. Fig. 8 shows some gain-frequency characteristics.

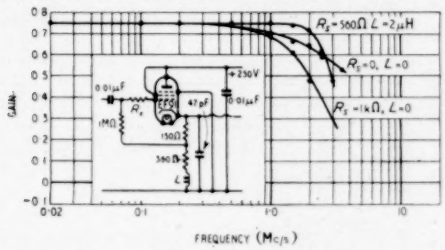

Fig. 8. *Effect of cathode inductance and grid stopper on gain.*

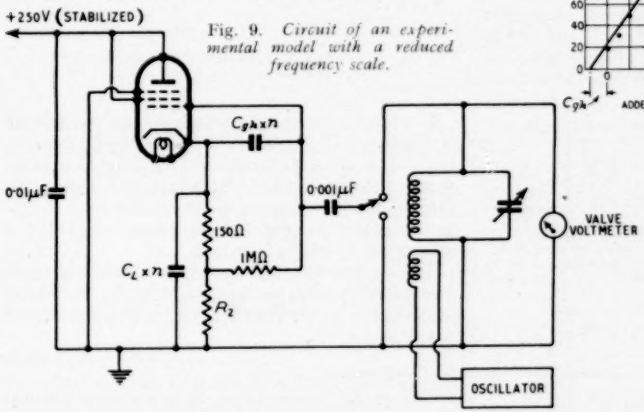

Fig. 9. *Circuit of an experimental model with a reduced frequency scale.*

In these experiments the measured input conductance differed from that calculated from equation (5) and, in particular, zero conductance occurred at a frequency of over one and a half times the frequency calculated from equation (7). For example, when $R_1 = 150$ ohms, $R_2 = 360$ ohms, $R_g = 1$ MΩ, $C_L = 53$ pF (47 pF + strays), $C_{gk} = 8$ pF and $g_m = 7{\cdot}5$ ma/V, the calculated frequency (f_o) at which the input conductance should change sign is 570 kc/s, whereas the measured frequency (f_o') was 890 kc/s; therefore, $f_o'/f_o = 1{\cdot}56$.

This discrepancy, which is due to the Miller effect, is explained in Section (4.3.). When $C_i = 7$ pF, $R_l = 2{\cdot}4$ MΩ and the input conductance is zero at 890 kc/s, the value of grid-stopping resistance required [from equation (25)] should be approximately 270 ohms. Fig. 7 shows that the grid-stopping resistance required in practice was 300 ohms.

4.2. *Experiments at Reduced Frequency Scale*

If C_{gk} and C_L are each multiplied by n and the resistances in the circuit are unchanged, the frequency at which the input conductance is zero is divided by n [see equation (7)]. The circuit shown in Fig. 9 was built in order to make measurements at lower frequencies to reduce errors due to stray circuit elements. The frequency of zero input conductance of the cathode-follower (f_o) was found by adjusting the oscillator frequency until the valve-voltmeter reading was the same whether or not the cathode-follower was connected across the coil (the circuit being tuned to resonance in each case). Figs. 10 to 12 show the results obtained. From equation (7), $1/f_o^2 \propto C_{gk}$ when $C_{gk} \ll g_m R_k C_L$; Fig. 10 shows $1/f_o^2$ plotted against the added grid-cathode capacitance, verifying this law. The straight line cuts the negative horizontal axis at about 8 pF which is the cathode-grid capacitance of the valve itself. Fig. 11 shows the variation of f_o with C_{gk} when C_{gk} is not small compared with $g_m R_k C_L$. It verifies that f_o is a minimum when $C_{gk} = \frac{1}{2} g_m R_k C_L$ [equation (8)] and tends towards infinity when $C_{gk} = g_m R_k C_L$, but all the observed frequencies are about 10% higher than the calculated values. Fig. 12 shows the variation of f_o with cathode load

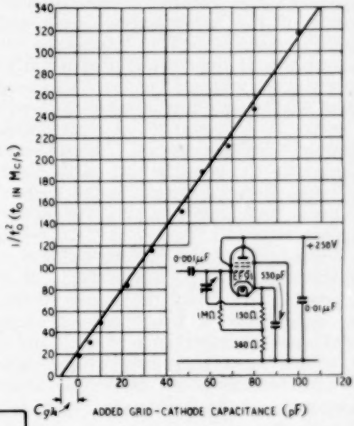

Fig. 10 (above). *Plot of added grid-cathode capacitance against* $1/f_o^2$.

resistance. Once again the observed frequencies are about 10% higher than the calculated values.

4.3. *Anode-Grid Feedback*

In the experiments on the scale model, the anode of the cathode-follower was connected to earth through the parallel combination of the impedance of the power-supply unit and the decoupling capacitor. The impedance of the power-supply unit (which included a hard-valve stabilizer circuit) was small at low frequencies but increased with frequency, becoming large compared with the reactance of the decoupling capacitor at high frequencies. Measurements of the voltage between anode and earth confirmed the net load. Because the valve had a capacitive anode load, feedback through the anode-grid capacitance increased the frequency at which the input conductance of the cathode-follower circuit passed through zero [equation (16)]. It is not easily possible to earth the anode through a smaller impedance: a 0·1 μF paper capacitor resonates with the inductance of itself and of the circuit wiring at a frequency of the order of 1 Mc/s and represents an increasing inductive reactance at higher frequencies.

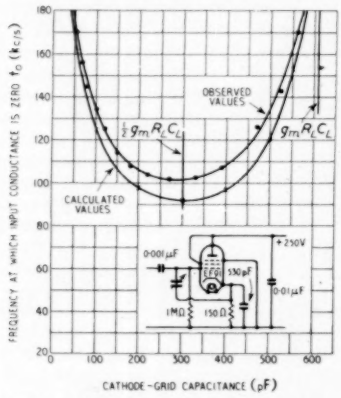

Fig. 11. *Variation of frequency of zero input conductance with cathode-grid capacitance.*

Since it was not possible to repeat the measurements of Section (4.2.) with the anode earthed much more effectively, a resistance (R_D) of 1 kΩ was connected in series with the anode of the cathode-follower and a capacitance C_A was connected between the anode and earth. Measurements were made of the frequency (f_o') for zero input conductance for different values of C_A; Fig. 13 shows the results obtained. Fairly good agreement is obtained between the measured curve and that calculated from equation (17) of Appendix 2; the discrepancy being due to the approximation made in Appendix 2. For example, f_o' should be a maximum when $\omega_o' C_A R_D = 1$ [equation (18)]: the observed maximum occurs when $C_A = 1200$ pF, $f_o' = 148$ kc/s; i.e., $\omega_o' C_A R_D = 1{\cdot}1$.

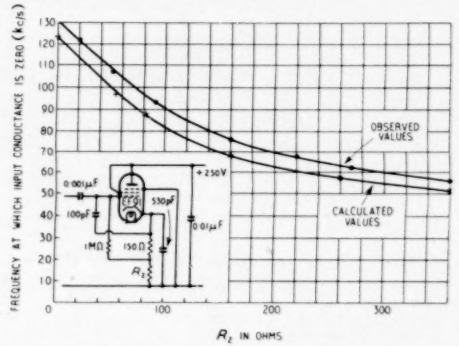

Fig. 12. *Variation of frequency of zero input conductance with cathode load resistance.*

Equation (16) should also hold for the results in Section (4.1.). When $C_A = 0{\cdot}01\,\mu$F, $f_o'/f_o = 1{\cdot}61$. The measured values of f_o in Section (4.1.) should, therefore, be 1·61 times the value calculated from equation (3), a ratio which is in fair agreement with the results obtained (see the example quoted).

The difference between the observed and calculated results in Section (4.1.) is, therefore, largely due to feedback through the anode-grid capacitance.

5. Conclusions

The input conductance of a cathode-follower with a capacitive cathode load is negative at all

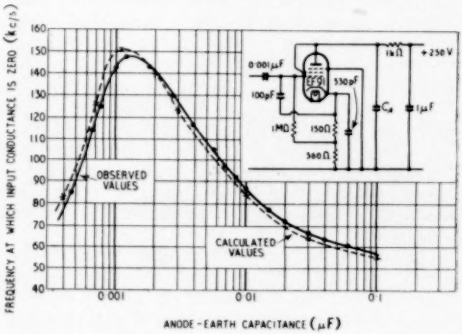

Fig. 13. *Variation of frequency of zero input conductance with anode-earth capacitance.*

frequencies greater than the critical frequency given by equation (7). When the anode of the cathode-follower is connected to earth by a finite decoupling capacitance, feedback through the anode-grid capacitance increases the critical frequency to the value given by equation (16), as was confirmed by experiment.

The only useful modification to the circuit which maintains a low positive input conductance at high frequencies is the addition of a series grid-stopping resistance. The value of grid-stopping resistance required is given by equation (25), which was confirmed experimentally. A triode valve is preferred to a pentode for a cathode-follower which is required to have a wide bandwidth and a low positive input conductance, because the grid-stopping resistance required by a pentode is usually so large as seriously to restrict the bandwidth of the circuit.

Acknowledgment

Acknowledgment is made to the Engineer-in-Chief of the G.P.O. and to the Controller of H.M. Stationery Office for permission to publish this paper.

APPENDIX 1

The Input Admittance of a Cathode Follower with a Capacitive Load.

The circuit is shown in Fig. 2 and its input impedance is equal to that of the equivalent circuit of Fig. 14. It is assumed that the valve is working on the linear part of its characteristic and that the current I_1 in the grid resistor is small compared with the cathode current.

$$V_{gk} = I_2 X_{gk} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

$$V_o = I_2 Z_k + g_m V_{gk} Z_k = I_2 Z_k (1 + g_m X_{gk}) \quad .. (2)$$

$$V_I = V_o + I_2 X_{gk} = I_2 (Z_k + g_m Z_k X_{gk} + X_{gk}) \quad .. \quad .. \quad .. \quad .. \quad (3)$$

where $Z_k = \dfrac{R_k}{1 + j\omega C_L R_k}$, $R_k = \dfrac{(R_1 + R_2) r_a}{R_1 + R_2 + r_a}$ and $X_{gK} = 1/j\omega C_{gk}$

$$\therefore \quad I_2 = \frac{V_I}{Z_k + g_m Z_k Z_{gk} + X_{gk}}$$

$$\text{But} \quad V_I = \frac{R_2}{R_1 + R_2} V_o + I_1 R_g \quad .. \quad .. \quad (4)$$

$$\therefore \quad I_1 = \frac{V_I}{R_g}\left[1 - \frac{Z_k R_2 (1 + g_m X_{gk})}{R_L (Z_k + g_m Z_k X_{gk} + X_{gk})}\right]$$

where $R_L = R_1 + R_2$

The total input current is

$$I_1 + I_2 + I_3 = \frac{V_I}{R_g} \cdot \frac{[(1 + g_m R_k) - j\omega(C_{gk} + C_L)R_k] \cdot [(1 + g_m R_1 - \omega^2 C_L C_{gk} R_g R_k) + j\omega(C_{gk} R_1 + C_L R_k + C_{gk} R_g)]}{(1 + g_m R_k)^2 + \omega^2 (C_{gk} + C_L)^2 R_k^2}$$

$$+ j\omega C_{ga} V_I$$

$$= V_I (G_i + j\omega C_i) \text{ where } G_i \text{ is the input conductance and } C_i \text{ is the input capacitance.}$$

$$\therefore G_i = \frac{(1 + g_m R_k)(1 + g_m R_1 - \omega^2 C_L C_{gk} R_g R_k) + \omega^2 R_k (C_{gk} + C_L)(C_{gk} R_1 + C_L R_k + C_{gk} R_g)}{R_g [(1 + g_m R_k)^2 + \omega^2 (C_{gk} + C_L)^2 R_k^2]}$$

$$\text{and } C_i = C_{ga} + \frac{(1 + g_m R_k)(C_{gk} R_1 + C_L R_k + C_{gk} R_g) - (C_{gk} + C_L) R_k (1 + g_m R_1 - \omega^2 C_L C_{gk} R_g R_k)}{R_g [(1 + g_m R_k)^2 + \omega^2 (C_{gk} + C_L)^2 R_k^2]}$$

But if R_g is large compared with R_L then

$$G_i = \frac{1}{R_g} \cdot \frac{(1 + g_m R_1)(1 + g_m R_k) - \omega^2 R_k R_g C_{gk} (g_m R_k C_L - C_{gk})}{(1 + g_m R_k)^2 + \omega^2 (C_{gk} + C_L)^2 R_k^2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

$$\text{and } C_i = C_{ga} + C_{gk} \frac{(1 + g_m R_k) + \omega^2 C_L (C_{gk} + C_L) R_k^2}{(1 + g_m R_k)^2 + \omega^2 (C_{gk} + C_L)^2 R_k^2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

The input conductance is zero at frequency $\omega_o/2\pi$ where

$$\omega_o^2 = \frac{(1 + g_m R_1)(1 + g_m R_k)}{R_g R_k C_{gk} (g_m R_k C_L - C_{gk})} \quad .. \quad .. \quad (7)$$

When $C_{gk} > g_m R_k C_L$, ω_o^2 is negative, therefore, ω_o is imaginary and when $C_{gk} = g_m R_k C_L$, ω_o is infinite. Therefore, the input conductance cannot become negative when $C_{gk} > g_m R_k C_L$. But ω_o is also infinite when C_{gk} is zero, so ω_o must be a minimum when C_{gk} has some value between zero and $g_m R_k C_L$. By putting $d(\omega_o^2)/dC_{gk} = 0$ it is found that ω_o^2 is a minimum when

$$C_{gk} = \tfrac{1}{2} g_m R_k C_L \quad .. \quad .. \quad .. \quad .. \quad (8)$$

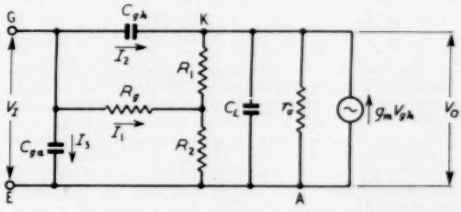

Fig. 14. *Equivalent circuit of Fig. 2.*

When $C_{gK} = g_m R_K C_L$, the gain of the circuit is independent of frequency, for, from equations (2) and (3), the gain is,

$$\frac{V_o}{V_I} = \frac{Z_k (1 + g_m X_{gk})}{Z_k + g_m Z_k X_{gk} + X_{gk}}$$

$$= \frac{g_m R_k}{1 + g_m R_k} \cdot \frac{1 + j\dfrac{\omega C_{gk}}{g_m}}{1 + j\dfrac{\omega R_k (C_L + C_{gk})}{1 + g_m R_k}}$$

$$= \frac{g_m R_k}{1 + g_m R_k} \quad \text{if } C_{gk} \text{ is put equal to } g_m R_k C_L \text{; i.e.,}$$

the gain is independent of frequency and the phase shift is zero.

But if, as is usual, $C_{gk} < g_m R_k C_L$, the gain has fallen by 3 db at the frequency $\omega_b/2\pi$ where

$$\omega_b^2 \left[\frac{(C_L + C_{ga})R_k}{1 + g_m R_k}\right]^2 - 2\left[\frac{C_{gk}}{g_m}\right]^2 = 1$$

$\therefore$ if $C_{gk}^2 \ll C_L^2$

$$\omega_b \approx \frac{1 + g_m R_k}{(C_L + C_{ga})R_k} \quad .. \quad .. \quad .. \quad (9)$$

The input resistance R_I of the circuit at low frequencies may be obtained by putting $\omega = 0$ in equation (5) and is

$$R_I = R_g \frac{1 + g_m R_k}{1 + g_m R_1} \quad .. \quad .. \quad .. \quad .. \quad (10)$$

i.e., the current has the same input resistance at low frequencies as a circuit with the same load resistance but with a grid resistance connected directly to the cathode, of value

$$R_{gk} = \frac{R_g}{1 + g_m R_1} \quad .. \quad .. \quad .. \quad .. \quad (11)$$

Substituting from equation (11) in equation (7) gives

$$\omega_o^2 = \frac{1 + g_m R_k}{R_{gk} R_k C_{gk}(g_m R_k C_L - C_{gk})} \quad .. \quad .. \quad (12)$$

APPENDIX 2

The Effect of Feedback Through the Anode-Grid Capacitance.

Because the anode of a cathode follower cannot usually be earthed except through a decoupling capacitance of finite susceptance, the results obtained in Appendix 1 will be modified by the Miller effect.

The input shunt resistance R_M of a valve due to the Miller effect is

$$R_M = \frac{-1/\omega C_{ga}}{|A| \sin\theta} \quad .. \quad .. \quad .. \quad .. \quad (13)$$

where A is the anode amplification of the stage and θ is the phase angle of the anode load impedance.

For a valve whose anode is connected to earth by a large decoupling capacitance C_A,

$$|A| \sin\theta = -g_m/\omega C_A$$

$$\therefore R_M = C_A/g_m C_{ga} \quad .. \quad .. \quad .. \quad .. \quad (14)$$

i.e., the component of the input conductance of a valve caused by a purely capacitive anode load is positive and independent of frequency.

The effect of feedback through the anode-grid capacitance upon the frequency at which the cathode follower has zero input conductance may be determined by substituting in equation (12) of Appendix 1 a new value for R_{gk} equal to the resistance of R_{gk} and R_M in parallel. Let this new value be R_{gk}'.

$$\therefore R_{gk}' = \frac{R_{gk} R_M}{R_{gk} + R_M} = \frac{R_g C_A}{g_m R_g C_{ga} + (1 + g_m R_1) C_A} \quad .. (15)$$

Therefore, a cathode follower with finite anode-earth susceptance has zero input conductance at frequency $\omega_o'/2\pi$ where

$$\omega_o' = \omega_o \sqrt{1 + \frac{g_m R_g C_{ga}}{(1 + g_m R_1) C_A}} \quad .. \quad .. \quad .. \quad (16)$$

where $\omega_o/2\pi$ is the frequency at which the input conductance would be zero if the anode were earthed through zero impedance [equation (7)].

The input conductance of the circuit is due to the component of the input current which is in phase with the input voltage, but the theory above assumes that the component of the input conductance which is caused by the Miller effect is due to the component of the input current which is in phase with the grid-cathode voltage. The grid-cathode voltage will not be exactly in phase with the input voltage because the phase angle of the cathode load is not zero; but the error is small in practice because the frequency of zero input conductance is usually small compared with the bandwidth of the circuit.

In the experiment described in Section (4.3), the anode-earth capacitances do not all have reactances small compared with the decoupling resistance R_D.

$$\therefore A = \frac{g_m R_D}{1 + j\omega C_A R_D}$$

$$\therefore |A| \sin\theta = -\frac{\omega g_m R_D^2 C_A}{1 + \omega^2 C_A^2 R_D^2}$$

$$\therefore R_M = \frac{1 + \omega^2 R_D^2 C_A^2}{g_m \omega^2 C_A C_{ga} R_D^2} \quad .. \quad .. \quad .. \quad .. \quad (17)$$

The value of C_A for which R_M is a minimum and ω_o' is a maximum is

$$C_A = 1/\omega R_D \quad .. \quad .. \quad .. \quad .. \quad (18)$$

when $R_M = 2/g_m \omega C_{ga} R_D \quad .. \quad .. \quad .. \quad (19)$

APPENDIX 3

Value of Grid-Stopping Resistance required to prevent Negative Input Conductance

The input admittance $(G_i + j\omega C_i)$ is given by equations (5) and (6). When $\omega > \omega_o$, the input impedance is equal to a resistance $-R$ in series with a capacitance and, if

$$-G_i \ll \omega C_i$$

$$-R \approx G_i/\omega^2 C_i^2 \quad .. \quad .. \quad .. \quad .. \quad (20)$$

Equations (5) and (6) may be rewritten as

$$G_i = \frac{C_{gk}(g_m R_k C_L - C_{gk})}{R_k(C_{gk} + C_L)^2} \cdot \frac{\omega_o^2 - \omega^2}{\omega_b^2 + \omega^2} \quad .. \quad (21)$$

$$C_i = C_{ga} + \frac{C_{gk} C_L}{C_{gk} + C_L} \cdot \frac{\omega^2 + \frac{\omega_b}{C_L R_k}}{\omega^2 + \omega_b^2} \quad .. \quad .. \quad (22)$$

where $\omega_o/2\pi$ is the frequency of zero input conductance [equation (7)] and $\omega_b/2\pi$ is the nominal bandwidth [equation (9)].

If $\omega \ll \omega_b$, $\frac{dC_i}{d(\omega^2)}$ is small and may be neglected.

$$\therefore \frac{dR}{d(\omega^2)} \approx \frac{1}{C_i^2} \cdot \frac{d}{d(\omega^2)}\left(\frac{G_i}{\omega^2}\right)$$

$$= \frac{C_{gk}(g_m R_k C_L - C_{gk})}{C_i^2 (C_{gk} + C_L)^2 R_k} \cdot \frac{d}{d(\omega^2)}\left(\frac{1}{\omega^2} \cdot \frac{\omega^2 - \omega_o^2}{\omega^2 + \omega_b^2}\right)$$

$$= 0 \text{ when } \omega^2 = \omega_o^2\left[1 \pm \sqrt{1 + \left(\frac{\omega_b}{\omega_o}\right)^2}\right]$$

If ω is real, $\omega^2 = \omega_o^2\left[1 + \sqrt{1 + \left(\frac{\omega_b}{\omega_o}\right)^2}\right]$ and corresponds to R passing through a maximum. But $\omega_o \ll \omega_b$

$$\therefore \omega^2 \approx \omega_o \omega_b \quad .. \quad .. \quad .. \quad .. \quad (23)$$

Substituting for ω^2 in equation (21),

$$G_i = -\frac{1 + g_m R_1}{R_g(1 + g_m R_k)} \cdot \frac{\frac{\omega_b}{\omega_o} - 1}{\frac{\omega_o}{\omega_b} + 1}$$

But $\omega_o \ll \omega_b$

$$\therefore G_i \approx -\frac{1}{R_I}\frac{\omega_b}{\omega_o} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (24)$$

where R_I is the input resistance at low frequencies [equation (10)]. R_S, the largest value of R, is obtained by substituting in equation (20) for G_i from equation (24) and for ω^2 from equation (23).

$$\therefore R_S \approx 1/\omega_o^2 C_i^2 R_I \quad .. \quad .. \quad .. \quad .. \quad .. \quad (25)$$

If a resistor is to be connected in series with the input terminals to prevent the input conductance becoming negative, its minimum value is given by equation (25). If $R_S \ll -1/G_i \ll R_g$ at $\omega = \sqrt{\omega_o \omega_b}$, it is immaterial whether R_S is connected in series with the input terminals or between R_S and the grid of the valve, the latter being the usual connection for a grid stopper.

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IMPEDANCE CHANGES IN IMAGE ICONOSCOPES

Methods of Automatic Compensation

By J. E. Cope and R. Theile, Dr.Phil.

(Pye, Ltd., Cambridge)

SUMMARY—This paper deals with the operation of the image-iconoscope type of television camera tube. To obtain a good signal-to-noise ratio a high value of load resistor must be used and, due to the capacitance across it, this causes signal integration. This must be compensated by a differentiation in the following amplifier. It was found experimentally that the required degree of compensation must be varied according to the average brightness of the televised scene, otherwise 'streaking' effects occur in the reproduced picture. Investigation showed that this was due to a change in the impedance of the camera tube. The varying impedance is a function of the secondary-emission current from the storage surface due to the changing stream of photo-electrons. Two methods of automatic compensation of this effect are discussed. It is shown that a suitable feedback to the signal plate provides an elegant and practical solution.

Introduction

IN television cameras, cathode-ray tubes with photo-sensitive surfaces are used.[1] These so-called pick-up tubes operate on the storage principle: the reflected light from the scene to be televised produces a corresponding charge pattern on a storage plate, either directly or by means of a photo-electron imaging process. The storage surface is periodically scanned by a suitably-deflected beam, thus translating the picture charge pattern into a succession of electrical impulses.

Depending on whether the arriving velocity of the scanning electrons is high or low, that is, on whether or not secondary-emission phenomena are present, we can divide the pick-up tubes into two groups. The first group (the high-velocity group) contains the iconoscope and Emitron, in which the photo-mosaic and storage surface are combined, and the image-iconoscope, super-Emitron and Photicon,* in which the homogeneous photo-cathode is separated from the storage surface by an image converter. The second group includes the orthicon and the CPS Emitron, which have a photo-mosaic similar to the normal iconoscope, and the highly sensitive image-orthicon which has an image converter and a secondary-emission multiplier for signal-current amplification.

The low-velocity tubes have been widely used because of their high sensitivity and relative absence of spurious signals. The high-velocity tubes, however, are still of considerable importance in television to-day. This is especially true of the image-iconoscope (Fig. 1) because of certain fundamental advantages, such as the comparative ease of manufacture, the stable operation at all light levels, the excellent half-tone rendering, and the fact that there are no serious limitations in achieving any reasonable degree of definition.

The advantage of being able to operate in a wide range of illumination can be, however, to some extent, lost in the connection between the pick-up tube and the amplifier. Without automatic compensation, 'streaking' effects in the

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* Trade mark of Cathodeon, Ltd., Cambridge, England.

picture become apparent when the illumination of the scene is changed by a comparatively large amount. The reason for this is a consequence of the use of a high value of tube load resistance R which is necessary in order to obtain the best signal/noise ratio.[2,3,4] This high resistance together with the tube, circuit and amplifier capacitances form a time constant which integrates the signal. The amplifier must, therefore, have a corresponding differentiating circuit to compensate this time constant, which has to be adjusted preferably by the use of a suitable test card; e.g., a black square on a white background.

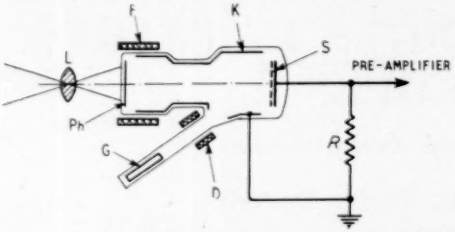


Fig. 1. *Schematic diagram of an image-iconoscope type tube (Photicon).*

Incorrect setting will result in 'trailing' or 'streaking' of the black square in the scanning direction, as illustrated in Fig. 2. The subjective effects of small distortions of this kind can be quite serious in the presence of vertically moving objects.

In the case of the Photicon it was found that the adjustment of this compensation was affected by the average brightness of the scene. When televising moving scenes with widely changing light levels, it proves impracticable for the operator to keep the compensation control accurately adjusted; therefore, automatic compensation is essential in order to make full use of the potentialities of this type of pick-up tube.

The following sections describe the investigations which led to an explanation of the changing input time constant and to a solution of the problem of automatic compensation.

Effect of Illumination on Photicon Impedance

A diagrammatic layout of a Photicon is shown in Fig. 1. Light from the scene to be televised is focused on to the photo-cathode by the lens L. The electrons released from the photo-cathode are focused on to the storage surface S by means of the magnetic lens F. The storage surface consists of a mica plate usually activated for a high secondary-emission ratio and backed with a conducting layer which forms the so-called signal plate. The surface S can be represented as a large number of discrete capacitances with a common electrode; i.e., the signal plate. The inside wall of the glass bulb is coated with a conducting film K which serves as an anode for the image converter and as a collector for the secondary electrons from the storage surface. An electron beam is produced by the gun G and deflected by the yoke D to scan the surface in the well-known manner.

It is not within the scope of this paper to discuss the mechanism of signal generation, but it is necessary to point out the various factors which influence the tube output impedance.

The velocity of the photo-electrons and the beam electrons is usually of the order of 1,000 V. The storage surface is bombarded with electrons from two sources: the photo-electrons in a continuous rain over the whole surface, the number of electrons per unit area depending on the scene to be televised, and the beam electrons concentrated into a spot, constant in number, and continuously sweeping over the storage surface. The bombarding electrons release secondary electrons at a rate, on the average, of three to five times the primary electrons.

Because the storage surface is highly insulating, the loss of secondary electrons makes it charge positively until an equilibrium potential is reached, at which point the retarding field so produced allows only the faster secondaries to reach the collector in numbers equal to the primary electrons. This equilibrium condition need only

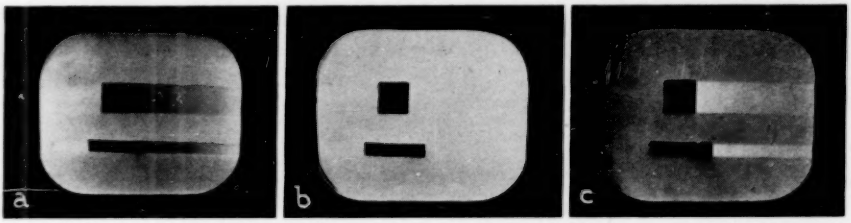


Fig. 2. *'Streaking' effects due to incorrect compensation for the amplifier input time constant; (a) under-compensated; (b) compensated; (c) over-compensated.*

be fulfilled over a complete scanning period, and the momentary current varies within this period according to the signal generation.

Fig. 3 illustrates the conditions existing at the storage surface, i_{pp} being the current from the photo-cathode, i_{pg} the current of the scanning

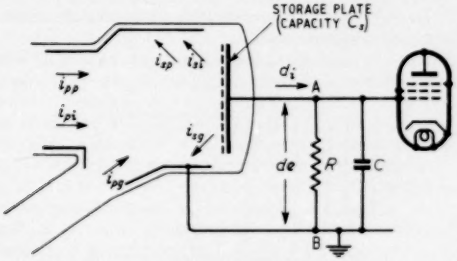
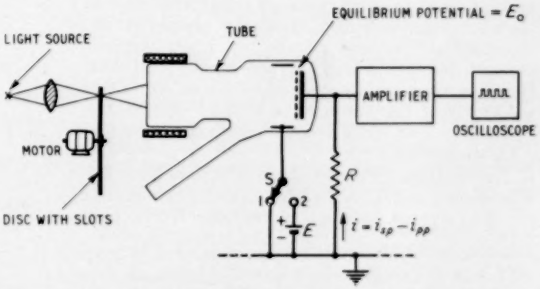


Fig. 3 (above). *Arriving and leaving currents at the storage surface.*

Fig. 4 (right). *A pulse method for the measurement of secondary-emission characteristics.*



electrons, and i_{sg}, i_{sp} the currents of the secondaries reaching the collector due to the scanning and photo-electrons respectively. In addition to these currents there will always be currents due to ionization because of the imperfect vacuum. Negative ions from the image-converter space are accelerated towards the storage surface while the positive ions are attracted to the photo-cathode and, due to the electrostatic lens field, are concentrated into a round patch on the photo-surface from which secondary electrons are released and imaged on to the storage surface in the same manner as the electrons released by photo-emission. The currents arising from the presence of ions are indicated by the arrows as i_{pi} and i_{si}.

It is now proposed to analyse the tube output impedance; that is, the impedance presented by the signal plate across the points AB in Fig. 3. It is apparent, because of the storage plate capacitance C_s, that to all intents and purposes there is no d.c. path and hence we are only concerned with the a.c. resistance or impedance de/di. This impedance could be shown as a series circuit comprising resistance resulting from secondary-emission effects, and capacitors which are part of the total storage plate capacitance. The variable element in this impedance is the quantity of photo-electrons, since the light falling on the photo-cathode changes with the scene to be televised. Thus the secondary emission characteristic due to photo-electron bombardment in conjunction with the capacitance of the bombarded area of the storage plate has to be analysed.

The time constant of this series circuit is of the order of 10^{-2} sec, if we take as a reasonable minimum only 10% illumination of the picture area, and if we estimate that the secondary-emission resistance is of the order of several megohms. On the other hand, the time constant of the tube-output/amplifier-input circuit is of the order of 10^{-4} sec, even if we assume fairly high values for the load resistance and total parallel capacitance (e.g., 5×10^{-5} sec for $R = 2 \times 10^{6}$ ohms and $C = 25 \times 10^{-12}$ farad). Therefore, the time constant of the photo-current circuit is long compared with that of the amplifier input circuit (i.e., the time constant to be compensated for); in other words, for the frequency range in which compensation takes place, the storage-plate capacitance can be regarded as a short circuit.

As a result of this reasoning, we can represent the effect of the photo-electrons as a resistance R_i in shunt with the input time constant.

It is important, therefore, to know how this resistance varies with average photo current. Since the photo-electron current is independent of the storage surface potential, R_i is given by the relation between the secondary-electron current and the potential difference between storage surface and collector.

It is not possible to measure this characteristic by normal d.c. methods, therefore, a pulse method is used.[5] The principle is shown in Fig. 4. The motor-driven light shutter produces pulses (at the rate of about 25 per sec) sufficiently short in duration for the charge on the storage-plate capacitance to be changed only a negligible amount, and thus this capacitance can, as previously suggested, be regarded as a short circuit. A variable voltage E is applied to the collector. At the instant of switching over to this voltage, a current flows through the load resistance R which is the difference between primary and secondary

electrons. The resulting voltage pulses, after suitable amplification, can be observed on the synchronously operated oscilloscope (utilizing a long-persistence screen powder). The oscilloscope is calibrated to read current.

With the switch S in position 1 the equilibrium potential E_o is established at the storage surface and no pulses are produced. By suddenly switching to the battery E (position 2) a current pulse flows through the load resistor R and the oscilloscope reading gives the signal current for the effective voltage $E - E_o$ (storage surface to collector) and thus a complete characteristic can be plotted by varying the battery voltage E, provided the equilibrium potential E_o is established by switching to position 1 between each step.

Measurements taken on a Photicon are plotted in Fig. 5. It was possible to present the various curves obtained for different values of photo-current i_{pp} in one average characteristic as shown. This indicates that the shape of the characteristic is not altered by photo-electron current density, which in turn means that only the law relating to the distribution of initial velocities of the released secondary electrons is applicable and space-charge effects are of negligible proportions.*

The resistance R_i is determined by the characteristic shown in Fig. 5. Within the operating range of the storage surface potential, R_i is substantially independent of E; i.e., constant. As the measurements result in one average characteristic, R_i is inversely proportional to i_{pp}.

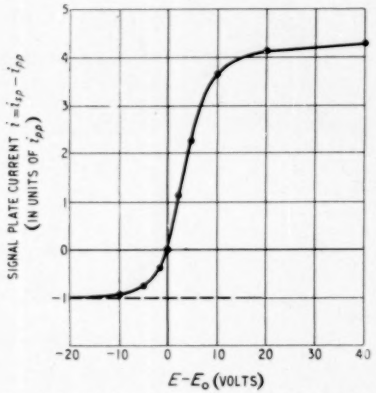


Fig. 5. *Secondary-emission characteristics for Photicon tubes.*

Hence

$$R_i = \frac{dE}{di_{pp}} = \frac{\text{const.}}{i_{pp}}$$

In the special case of the Photicon the constant is about 2 (ohm-amp).

$R_i = 2/i_{pp}$ (megohms, when i_{pp} is in μA). This relationship is shown in Fig. 6.

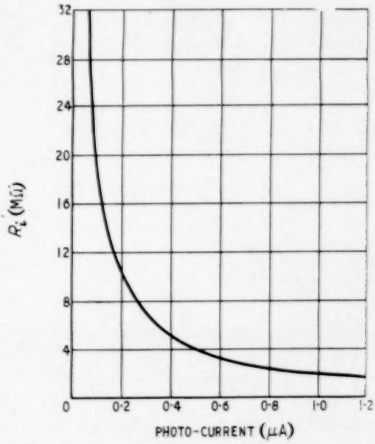


Fig. 6. *Graph showing the impedance* $R_i = 2/i_{pp}$.

The measurements in Fig. 5 were carried out with an idealized condition of even illumination over the whole photo-cathode surface, but further checks showed that, provided the average illumination was constant, the distribution of light and shade was only of second-order importance; i.e., the derived results are valid for any light distribution in a normal televised scene.

The tube resistance will be influenced by ion and electron currents caused by residual gases.

The effect of the scanning beam on the tube impedance is different because of the higher current density and the small part of the storage-plate capacitance which is in operation at any instant. It is not necessary to analyse this aspect, however, as the beam current remains constant in operation and in any case this part of the total impedance is not likely to change with changes of illumination.

Finally, a proof of the aforementioned measurements is given in Fig. 7. Here is shown the result of measurements made with several Photicon camera tubes under normal operating conditions. The compensating time constant in the amplifier

* It will be interesting to recall the relationship for the impedance r of a diode (thermionic emission). In the range where the transit of electrons from cathode to anode is determined by the initial velocity distribution the following relationship holds:

$r = c_o/i_p$ where c_o = the average initial velocity measured in volts.

i_p = cathode current.

On the other hand in the space-charge region (i.e., the normal operating region for thermionic tubes)

$$r = \text{const.}/\sqrt[3]{i_p}$$

was adjusted and measured in each case and by means of a simple calculation it was possible to work backwards and isolate the varying part of the tube resistance. The mean curve in Fig. 7 can be seen to approximate closely to the curve in Fig. 6. Fig. 7 also indicates the tolerance to be expected due to imperfect vacuum, variation in secondary-emission ratios, etc.

Summarizing the results of these investigations we may say, as shown in Fig. 8, that the tube-output to amplifier-input circuit comprises the load resistance R, which includes the constant part of the tube output resistance, the variable element R_i, which is in inverse proportion to the photo-electron current, and the capacitance C which is the sum of the tube, wiring and amplifier input capacitance, all in shunt. While Fig. 8 is not a general statement of the tube output circuit, it is sufficiently complete for the problem discussed here.

Automatic Compensation Methods

1. *Use of Low Values for R*

According to the previously mentioned investigations the range of variation of R_i under normal operating conditions (i.e., photo-currents of the order of 0·1 to 1·0 μA) is between 20 and 2 megohms.

The relative changes of the input time constant

$$\tau = \frac{RR_i}{R + R_i} C$$

could be small enough if

$$R \ll R_i$$

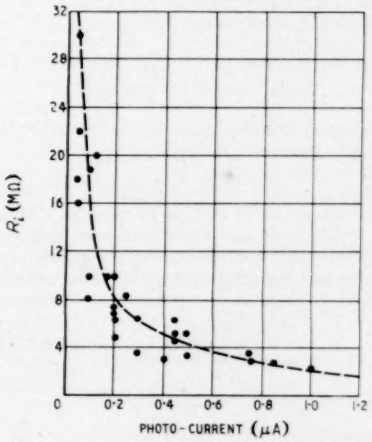


Fig. 7. *The result of impedance measurements on several Photicon tubes.*

On the other hand, a high value for the load resistance R is most desirable not only because of the improved signal/noise ratio, but also in order to obtain high amplitude in the low-frequency region to overcome the effect of amplifier microphony, hum, etc.

Thus it is obvious that the elimination of streaking effects by use of a low value of R means a sacrifice of considerable advantages.

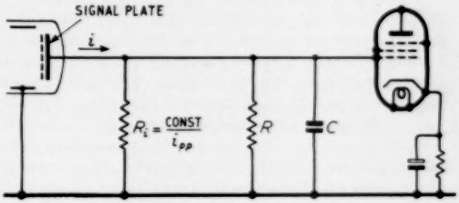


Fig. 8. *Tube-output to amplifier-input circuit.*

2. *Automatic Control of the Compensating Time Constant of the Amplifier*

A straightforward solution for automatic compensation is shown in Fig. 9. It is especially applicable to the type of compensation circuit which has the compensating time constant as a feedback network in the cathode circuit of an amplifying stage B.[6]

For compensation, the time constant $R_K C_K$ must equal the input time constant. In parallel with R_K there is the resistance r_a of a triode. The photo-cell output current develops a potential, across R_L which is applied through a filter, $R_f C_f$ to the control grid. The photo-cell is illuminated by a part of the light flux, which is focused on to the photo-cathode of the pick-up tube (by means of a small semi-transparent mirror, by reflection, etc.).

By selection of a suitable range of the triode characteristic, it is possible to get an equal change of both time constants and, therefore, a compensation which holds over a wide range is obtained.

It is also possible to derive the control voltage directly from the photo-electron current of the Photicon. Other possibilities for controlling the time constant, according to the amount of light falling on a photo-cell, will occur to the reader.

Experiments with circuits as shown in Fig. 9 showed good results. This method, however, is relatively complicated in comparison with the feedback method described later, and certain disadvantages are involved. For example, any changes in the compensating circuit directly influence the accuracy of the automatic control. Also, only one special impedance characteristic is compensated. Thus, if the pick-up tube is changed, a new adjustment may have to be made.

3. *Automatic Compensation by Feedback to the Signal Plate*

The application of feedback to the signal plate was found to be a much more elegant solution, because full compensation is obtained for any change of the tube resistance R_i over a wide range.

Fig. 10(a) illustrates the principle. The signal plate is connected to the input of the pre-amplifier. This amplifier has a voltage gain A, and the output voltage e_o is in antiphase to the input voltage e_i (odd number of stages). The load resistor R is connected to the amplifier output, giving negative voltage feedback to the signal plate. At the amplifier input is the impedance Z (tube impedance in shunt with input capacitances).

By examining the relationship between the signal current and the pre-amplifier output voltage e_o, a better understanding of this principle will be obtained.

The amplifier with feedback has a voltage gain of:

$$\frac{e_o}{e_i} = \frac{1}{\frac{1}{A} + \frac{Z}{Z+R}}$$

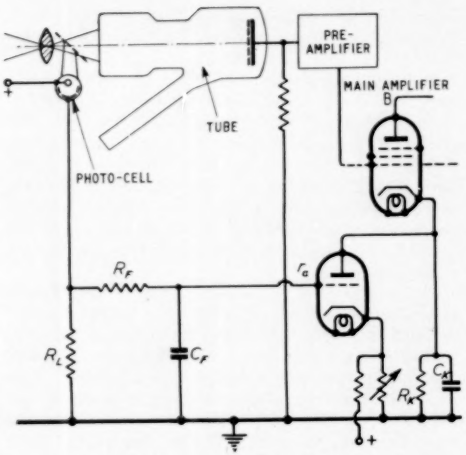


Fig. 9. *Compensation of tube impedance changes by automatic control of the correcting time constant in the amplifier.*

The Photicon generates current pulses i, which represent the television signal. The signal voltage e_i equals this current i, multiplied by the resultant impedance of the signal circuit.

$$e_i = i\,\frac{RZ}{R+Z}$$

It is assumed that the amplifier output impedance is small compared with R and that the source impedance is sufficiently high for the effect of the signal circuit on i to be negligible. Experimental evidence confirms this assumption.

The output voltage is then:—

$$e_o = i\,\frac{RZ}{R+Z}\,\frac{1}{\frac{1}{A} + \frac{Z}{Z+R}} = iR\,\frac{A}{\frac{R}{Z} + 1 + A} \quad \ldots (1)$$

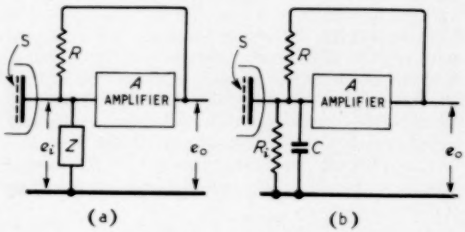


Fig. 10. *Principle of feedback to the signal plate.*

It is apparent from equation (1) that if $(1 + A)$ is large compared with R/Z then e_o will be independent of the impedance Z. Therefore, by a suitable choice of the amplification A any desired degree of immunity from impedance changes can be obtained.

The feedback circuit is now considered for the special case of Fig. 10(b). We have to write for Z:

$$Z = \frac{R_i}{1 + pR_iC} \text{ with } p = j\omega$$

Equation (1) then changes to

$$e_o = i \cdot a\,\frac{1}{1 + p\,\frac{RC}{1 + A + R/R_i}} \quad \ldots \quad \ldots \quad (2)$$

in which

$$a = \frac{AR}{1 + A + R/R_i}$$

This result gives an idea of the mode of operation of the feedback-compensation principle. It is obvious that the frequency response of the signal is no longer determined by the time constant

$$\tau = \frac{RR_i}{R + R_i}\,C$$

but by the smaller time constant,

$$\tau' = \frac{RC}{1 + A + R/R_i} \approx \frac{RC}{1 + A} \quad \ldots \quad \ldots \quad (3)$$

in which R_i is of much lower influence. This can also be explained as follows: in a system with a large amount of feedback, the amplification is

given mainly by the feedback ratio. This ratio is in our case the potential-divider ratio:

$$1 + \frac{R}{Z} = 1 + \frac{R}{R_i}(1 + pR_iC)$$

The time constant of the input circuit influences this ratio and the signal amplitude in such a manner that these two effects nearly compensate each other.

Owing to the fact that the feedback decreases with increasing frequency, this compensation is not effective for an unlimited frequency range. Some compensation at the upper end of the video band will be necessary due to the need for keeping the gain A to a practical value, but in this frequency range the reactance of the capacitance C is numerically much smaller than R_i and, therefore, changes in R_i are of no importance.

The following estimation shows that a relatively small value for A is sufficient to guarantee the required compensation. We assume now that the load resistor is 2 megohms and that the tube resistance changes from infinity down to about 4 megohms under normal conditions. Thus the maximum value for R/R_i will be 0·5 and it follows from equations (2) and (3) that if A equals 50, any change of R_i within the operating range will modify τ' by less than 1%. Experiments confirmed that with this value of A all streaking effects disappeared completely.

Compensation by the use of feedback to the signal plate is of special advantage. Not only are the effects of the tube resistance change due to varying photo-current (scene brightness) reduced, but any streaking due to other reasons (such as variation of the beam current and second-order effects which have not been discussed in this paper) is minimized. Another advantage of this method of compensation is that it is not necessary to know the impedance function of the various parameters.

The feedback to the signal plate according to Fig. 10 also affects the necessary differentiation of the signal, and it is for this purpose pre-amplifier systems of this kind have been proposed before and are currently used.[7,8,9] We shall now consider this aspect of the circuit operation. A distortionless picture signal requires that the amplifier output voltage e_o be proportional to the signal current i in respect to both frequency and phase.

Equation (2) shows that this is fulfilled if

$$\omega \frac{RC}{1 + A + R/R_i} \ll 1 \qquad .. \quad .. \quad (4)$$

which in turn means that A must be sufficiently large.

The frequency at which the signal power drops to one half (or the voltage to 0·7) can be defined as follows:

$$\omega_m = \frac{1}{\tau'} = \frac{1 + A + R/R_i}{RC}$$

or

$$A \approx \omega_m RC$$

thus when

$C = 20$ pF; $R = 2 \times 10^6$ ohms; $RC = 4 \times 10^{-5}$sec

then for a half-power frequency $f_m = 5$ Mc/s an amplification of approximately 1,200 is needed. Such a high value of amplification in the feedback loop is inconvenient, because of the difficulties with instability outside the used bandwidth. Because (4) depends on the frequency, these difficulties may be considerably reduced in the lower-frequency range by a simultaneous frequency-dependent decrease of the amplification A down to a minimum of about 50 (to guarantee fully the streaking compensation). It is plain that for the lower frequencies quite a low amplification is sufficient to fulfil the conditions in equation (4). The function $A(\omega)$ may deviate considerably from a linear characteristic provided equation (4) is satisfied—in other words, the amplification A must be sufficient to guarantee that the overall response is mainly dependent on the feedback ratio and not on the normal response curve of the amplifier.

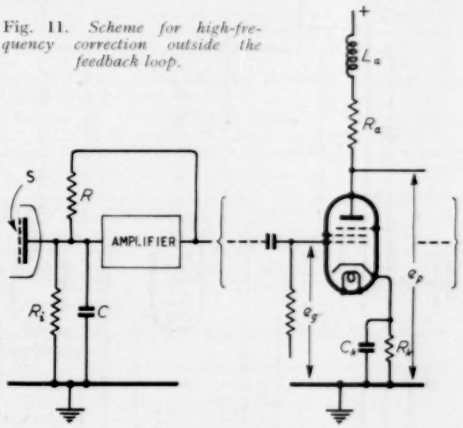


Fig. 11. *Scheme for high-frequency correction outside the feedback loop.*

Apart from this method of complete frequency compensation (differentiation) of the television signal within the feedback loop, there is the other possibility of additional compensation in the following main amplifier. In this case A is constant and as large as necessary to get complete streaking compensation (i.e., $A = 50$).

According to (3) the following amplifier must

compensate for the time constant:

$$\tau' = \frac{RC}{1+A} \; ; \; 1 + A \gg \frac{R}{R_i}$$

The amplifier characteristic must then be of the form:

$$c'\left(1 + p\,\frac{RC}{1+A}\right)$$

where c' = constant

which can be obtained in various ways. Fig. 11 shows an example relating to the known method with current feedback in the cathode circuit of a valve.

A simple calculation shows the characteristic $G(\omega)$ of such an amplifier[6]

$$G(\omega) = \frac{e_p}{e_g} = g_m R_a \frac{(1 + pR_K C_K)(1 + pL_a/R_a)}{1 + g_m R_K + pR_K C_K}$$

g_m = mutual conductance of the amplifier valve. The total amplification of the combined amplifier system gives

$$e_p = e_o . G = i\,\frac{a g_m R_a}{1 + g_m R_K} \quad \frac{(1 + pL_a/R_a)(1 + pC_K R_K)}{\left(1 + p\,\frac{RC}{1+A}\right)\left(1 + p\,\frac{R_K C_K}{1 + g_m R_K}\right)}$$

The frequency-conscious part falls out if the conditions

$$\frac{RC}{1+A} = R_K C_K \text{ and } \frac{R_K C_K}{1 + g_m R_K} = \frac{L_a}{R_a}$$

are fulfilled. In that case the output voltage is really proportional to the signal current.

According to the tolerances, the two time constants

$$\tau_K = R_K C_K \text{ and } \tau_l = L_a/R_a$$

may have to be adjusted.

The decision as to which frequency-response compensation (differentiation) should be used depends on several practical points of view.

The complete compensation in the head amplifier is the ideal system. The signal leaves the camera corrected and there is no need for readjustment if the camera tube is changed. On the other hand, the high value of amplification required makes the amplifier more complicated, especially as the frequency response of the feedback loop has to be considered beyond the used bandwidth to avoid instability. Also the total

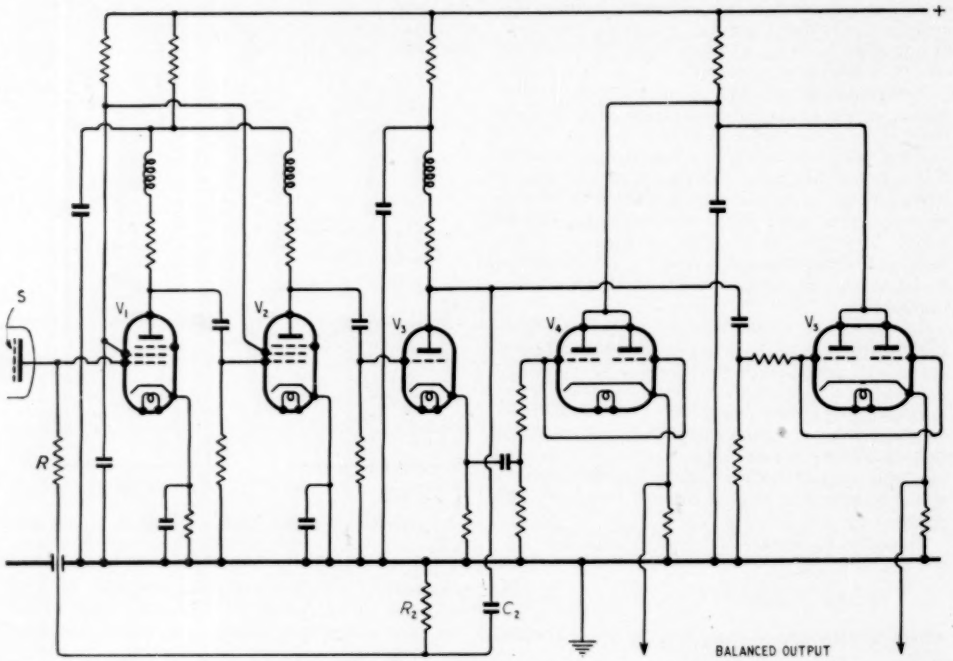


Fig. 12. *A pre-amplifier circuit with feedback to the signal plate to suppress streaking due to tube impedance changes.*

delay inside the loop must be very small. A major difficulty is the effect of stray capacitance across the load resistor and capacitances from this resistor to earth, but a subdivision of this resistor minimizes these effects considerably.[7] None of these difficulties exists when using the second method; i.e., with the compensation in the high-frequency region outside the feedback loop. The pre-amplifier is thus simplified and is more adaptable for fitting in a space limited camera. A disadvantage, however, is the necessary adjustment of the compensating time constant in the following amplifier.

Fig. 12 shows the circuit of a three-stage pre-amplifier for the second method. The feedback is ineffective in the range of high frequencies where appreciable phase distortion may occur. In the low-frequency region special care must be taken in order to avoid regeneration. This is achieved by ensuring that the amplification A falls off linearly with decreasing frequency down to and below the frequency where A is less than unity, which in turn is achieved by inserting a small coupling time constant in the amplifier, located between the anode of the second valve and the grid of the third stage, while all other time constants are greater by some 20 times.

The described principle of streaking compensation has no disadvantages in other respects. The signal-to-noise ratio is not deteriorated. For sufficiently high values of R and a fixed bandwidth, this ratio depends only on the total input capacitance and the equivalent noise resistance of the first amplifier valve. It is immaterial whether the compensation is done by feedback or by other methods elsewhere in the amplifier chain.[10]

The compensation of streaking with feedback has been applied in practice for some time now, those cameras in use with the B.B.C., made by Pye, Ltd., Cambridge, employ this principle, which has been found satisfactory in operation.

Acknowledgment

This work was carried out in the laboratories of Pye, Ltd., and Cathodeon, Ltd., both of Cambridge, to whose directors we are indebted for permission to publish this paper. The authors wish to express appreciation for the encouragement given by the technical director, B. J. Edwards, M.B.E., M.I.E.E. Thanks are due to K. Whiting and H. A. McGhee, Grad.I.E.E., for their help in carrying out the experiments which led to the publishing of this paper.

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PATH-LENGTH MICROWAVE LENS

Experimental Characteristics

By D. G. Kiely, M.Sc., A.Inst.P.

(Royal Naval Scientific Service)

SUMMARY.—An experimental investigation of the E-plane radiation patterns of a path-length lens at 3-cm wavelength is described. It is shown that the lens introduces a considerable phase curvature in the E-plane which is highly detrimental to the radiation pattern. An estimate of the magnitude and an explanation of the effect are given.

Introduction

PATH-LENGTH microwave lenses were first introduced by Kock[1] in 1949. The lens is of the delay type and consists of a set of parallel metal plates in planes parallel to that of the magnetic vector of the incident electromagnetic wave. The plates are inclined at an angle θ to the axis of the aerial system and lie within a contour which is derived by the normal methods assuming a known dielectric constant which is greater than unity. (See Fig. 1.) The wave is constrained to the channels between the plates, but the wave phase velocity is not changed as the plates are parallel to the magnetic vector. Phase correction, that is, the production of a uniform phase distribution across the lens aperture, is achieved by varying the actual length of the ray path within the lens. It can be readily seen that the refractive index of the lens medium is equal to secant θ. In his paper Kock states that because of the shift in energy distribution in the electric plane as the

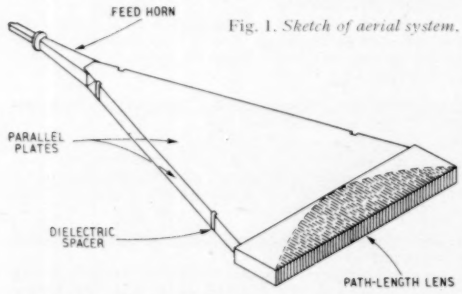

Fig. 1. *Sketch of aerial system.*

waves pass through the tilted plates of the lens, the minor lobes are not as well suppressed as in the magnetic plane. The present experimental investigation of E-plane radiation patterns of a cylindrical path-length lens has shown that the effect of the lens on the patterns is much more serious than is implied by Kock and that, in fact, an asymmetric phase curvature is introduced across the aperture by the lens; that is, the phase distribution across the lens aperture, which should be constant, is a curve which does not possess axial symmetry. This produces an asymmetric main beam which could not be caused by an asymmetric distribution of field *amplitude* alone. It is to be expected that both the phase and the amplitude distributions associated with a plano-convex lens will be asymmetric and the investigation described in this paper gives an estimate of the magnitude of the effect on the radiation pattern of a typical 4-ft aperture cylindrical lens designed for 3·16-cm wavelength.

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Experimental Procedure

A plano-convex cylindrical lens was constructed with the angle θ of 50° between the lens plates and the axis giving a refractive index μ of 1·56 according to the relation for the path-length lens.

$$\mu = \text{secant } \theta \quad .. \quad .. \quad .. \quad (1)$$

The lens aperture and the focal length were 4 ft and the lens height was made 3 in. A pair of metal parallel plates extended from the lens aperture to the feed horn at the focus. The feed horn was a small electromagnetic horn of aperture 3 in arranged to radiate a horizontally-polarized wave. The aerial structure is shown in Fig. 1. The sides of the parallel plates were left open as, with horizontal polarization, excessive wall reflections leading to many large sidelobes occur if the sides of the parallel plates are closed with metal. The spacing between the plates of the lens was made 0·32 in to ensure that no higher mode wave propagation occurred between the lens plates.

The horn was placed with its aperture at the focus and its axis coincident with the lens axis. In this position the aperture distribution of field amplitude is asymmetric due to the action of the tilted plates on the symmetric diagram of the feed horn, also it is to be expected that a certain amount of asymmetric phase curvature will be introduced by the lens itself as the angle between the lens plates and the incident ray is not constant across the aperture and a phase change will arise at this discontinuity in the ray path. This second asymmetry will produce an asymmetric main

beam and sidelobe structure. The radiation pattern of the lens with the horn symmetrically placed at the focus was recorded and is shown in Fig. 2(a). The pattern is quite asymmetric, indicating considerable phase unbalance across

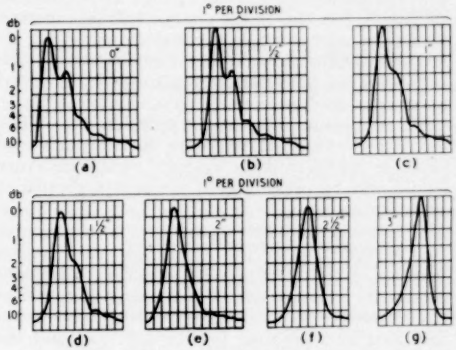

Fig. 2. *Radiation pattern of lens with the horn symmetrically placed at the focus (a) and with various displacements from the axis (b-g).*

the aperture. To correct this the feed horn was moved laterally off the axis in $\frac{1}{2}$-in steps in the direction of decreasing angle between lens plates and incident ray. Radiation patterns were recorded for each $\frac{1}{2}$-in position for a total displacement of 3 in and are shown in Fig. 2 (b, c, d, e, f and g). It will be noticed that the asymmetry gradually decreases until at $2\frac{1}{2}$ in the main beam is reasonably symmetric; at 3 in asymmetry in the opposite sense is present.

The diagram corresponding to $2\frac{1}{2}$-in displacement must be associated with a symmetric (not necessarily uniform) phase distribution across the aperture so it is reasonable to assume that the phase curvature due to the lens is here compensated by the phase curvature due to the feed displacement. It is assumed that the inherent

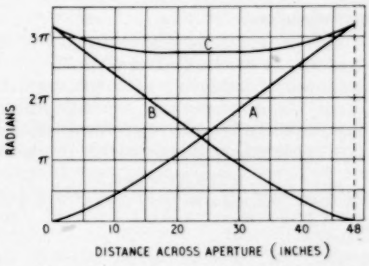

Fig. 3. *Phase curvature, A due to feed displacement, B to lens and C the resultant.*

phase curvature of the lens is unaffected by the feed displacement of $2\frac{1}{2}$ in which is a reasonable assumption as the range of angles between incident rays and lens axis is only slightly altered by the feed displacement. The feed-displacement curvature may be calculated from the geometry of the lens and is shown in Curve A, Fig. 3. The complementary curve of this, Curve B, is then taken as representative of the phase curvature due to the lens, and the addition of these two phase curves produces Curve C, which is practically uniform and is associated with the radiation pattern of Fig. 2(f). This phase curve departs from uniformity only at the edges and as the field strength there is low using a 3-in aperture feed horn, the effect on the radiation pattern is small. An approximation of the amplitude distribution may be obtained from the simple geometry of the lens knowing the primary-feed radiation pattern and making the assumption that the reflection coefficient of the lens is uniform across its aperture. This is shown in Fig. 4.

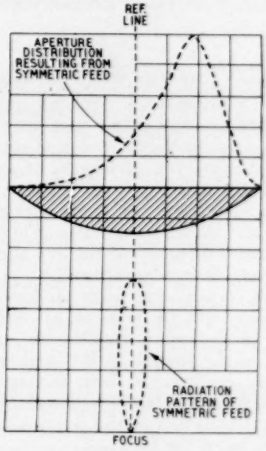

Fig. 4. *Approximation to the amplitude distribution across the lens.*

Curve B, Fig. 2, shows that the asymmetric phase curvature caused by the lens reaches a value of $3{\cdot}2\pi$ and so the phase-correcting function of such a simple plano-convex lens is so seriously impaired by this effect that the lens, for E-plane operation, cannot be designed in the straightforward manner normally employed for plano-convex delay lenses. The inherent phase curvature of the lens must be considered in the design which will lead to a displacement of the feed from the focus, or an orientation of the feed off the lens axis and, although this may be acceptable for E-plane diagrams only, it will not produce

focused beams in both E- and H-planes. Thus the principles of operation and design of the path-length lens are not quite so simple as is commonly believed.

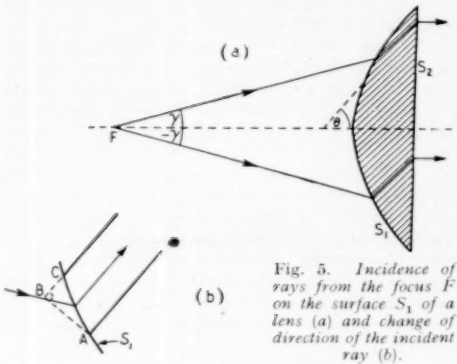

Fig. 5. *Incidence of rays from the focus F on the surface S_1 of a lens (a) and change of direction of the incident ray (b).*

Inherent Phase Curvature of a Path-length Lens

To obtain a general idea of the important factors governing the inherent phase curvature of a path-length lens consider the incidence of rays from the focus F in the convex surface S_1 of a lens, Fig. 5(a). If θ is the angle between the lens plates and the lens axis and γ is the angle between the incident ray and the lens axis, then [Fig 5(b)] when the ray changes direction at the lens surface it will suffer a phase change which will be proportional to the angle $(\theta - \gamma)$ and when $(\theta - \gamma)$ is zero the phase change will be zero or at least a small minimum. Also, due to the field concentration in the region of the triangle A B C and in the space immediately outside the lens surface the phase change will be proportional to the plate spacing, t. So, an approximation to the relationship may be assumed to be of the form

$$\Delta\phi = Kt(\theta - \gamma)^n \quad .. \quad .. \quad .. \quad (2)$$

Where $\Delta\phi$ = phase change at lens surface S_1
and K = a constant

Re-plotting Curve B, Fig. 3, as a curve of $\Delta\phi$ against $(\theta - \gamma)$ on logarithmic paper a straight line is obtained which yields the values $n = 3{\cdot}85$ and $K = 1{\cdot}96 \times 10^{-6}$.

So equation (2) for the aperture plano-convex lens with plate spacing 0·32 in becomes

$$\Delta\phi = 1{\cdot}96 \times 10^{-6}t(\theta - \gamma)^{3{\cdot}85} \quad .. \quad (3)$$

with $\Delta\phi$ in radians
t in inches
θ and γ in degrees

There will also be a phase change at the second surface S_2 but this phase change will be constant across the aperture as γ is constant across S_2.

The asymmetric aperture distribution of the plano-convex path-length lens may be improved by using a double-convex lens in which the plates slope in opposite directions in each half. The method of lens design to obtain the contour would be quite normal but the inherent phase curvature would still be present with the feed horn placed symmetrically on the axis at the focus. The use of a double-convex lens will not completely remove the asymmetry in the amplitude distribution because of the variation in reflection coefficient across the lens surface. This is also a function of the difference in the angles between lens plates and axis and between the incident ray and the axis and tends to give an asymmetry in the same sense as that produced by the tilt of the plates.

Conclusion

The above considerations do not apply to the H-plane where the aperture phase and amplitude distributions are uniform and the radiation patterns correspondingly symmetric. For E-plane applications which require a symmetrical beam and low sidelobes the plano-convex lens cannot be satisfactorily designed in the normal straightforward manner. The inherent phase curvature of the lens must be taken into consideration which will involve either a more complicated mathematical analysis in determining the lens contour or an empirical method of fixing the position of the feed horn.

With these considerations it is concluded that the path-length lens, although having many advantages for H-plane applications, is fundamentally unsuited for phase correction in the E-plane.

Acknowledgments

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WAVEFORM SYSTEMS AND 'TIME EQUALIZERS'

Their Transmission Characteristics

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(Communication from the Staff of the Research Laboratories of The General Electric Co., Ltd., Wembley, England.)

SUMMARY.—This paper describes a new approach to the solution of several of the outstanding problems in circuit theory where the available transmission information has been given in terms of distorted waveforms. It is shown that some of the classical transform concepts relating 'frequency' and 'time' characteristics can be replaced by much more straightforward algebra in which the physical interpretations are constantly in view. The amplitude and phase characteristics of a system, the internal structure of which can be unknown, are evaluated quite simply from the non-redundant networks which are arranged to simulate the actual system waveform distortions. The insertion characteristics of the waveform-correcting networks, or 'time equalizers', follow directly from this first step.

Extensions of the method give practical solutions for the waveform properties of any number of systems working in tandem, or for any fraction of a known system. In these cases it has been found unnecessary to revert to frequency characteristics as the manipulations are carried out directly in terms of the echoes, or individual parts of the waveform distortions.

In conclusion, it is suggested that there are appropriate roles for 'time equalizers' and the better known 'steady-state' equalizers, but the correction of some systems may require the use of both types of networks.

1. Introduction

MANY of the difficulties experienced in the study of circuit distortion problems arise from the emphasis placed on the calculation of the transmission characteristics from a knowledge of the actual system structure. Calculations of this kind are useful in the synthesis of small parts of a system and an approximate evaluation of the total distortion, in steady-state terms, can be obtained by the addition of the separate parts. Even so, the errors from interaction and reflection factors between the many parts are difficult to determine in a typical extensive system, such as a radio link with repeaters. The problems are so great in such cases that any attempts at a solution of the distortion problem would be confined to the amplitude/frequency characteristic alone, and the residual over-all distortion to be corrected would be found by steady-state measurements of amplitude characteristics, and also by the measurement of appropriate phase parameters in the cases of systems defined by waveform quality.

The treatment of a system by the time characteristics of parts or the whole would lead to an untenable position, even in the early stages. The calculation of time characteristics from the differential equations of an intervalve coupling network comprising only 10 elements presents enough difficulties to deter the engineer from attempts to enquire any further. There is a very small number of exceptions; for example, a smooth cable with primary constants per unit length unchanged over the whole length can be treated equally well by both steady-state and transient methods. However, the case is of little practical interest as the distortion could be corrected by 'steady-state' equalizers alone, but the transmission of waveforms in practice is dominated by the macroscopic effects of distributed irregularities which require time equalizers for correction.

It is suggested here that the use of the differential equation as a definition of the transfer constant of a complex system is almost useless in most cases. The starting point in the procedure involving time characteristics should be the information available at the far terminals when an appropriate stimulus is applied at the near terminals. This approach can be developed to yield several important results of practical interest without reference to the circuit structure of the system. It must be accepted that, at this date, any correction of time or frequency characteristics is valid only if the system is linear.

Some of the problems in the transient state have been solved hitherto in neat symbolic forms which, in the author's opinion, are misleadingly simple as they usually mask the extreme labour required to achieve any practical or numerical results. The present need is to make better use of the fundamentally reciprocal nature of frequency and time.

In a previous paper[1] the author has shown that a transmission system giving rise to distortion of echo form can be represented by the impedance

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characteristics of a two-terminal simulating network. The elements of this hypothetical network can be determined almost by inspection from the actual distorted waveform, and the structure of the network can be arranged in the best form for the calculation of the simulating impedance.

Evaluation of the simulating impedance itself may be considered mainly of academic interest, but the inverse of this impedance defines the steady-state properties of the practical time equalizers for which it is important to know the insertion loss. Furthermore, the simulating impedance gives, by a calculation well understood by circuit engineers, the transform of the system waveform function without any reference to the less direct concept of the Fourier integral.

2. Impedance Transform Characteristics

It is clear that the various series and cascade forms of simulating networks appropriate in the case of a given system distortion must be exactly similar in impedance characteristics. However, the networks do not present the same ease in the calculation of impedance and the Series Type (2) of Fig. 1 is chosen as the most straightforward. In this network the lines have echo times $t_1 t_2 t_3 \ldots t_n$ and the electrical 'lengths' $\theta_1 \theta_2 \theta_3 \ldots \theta_n$ are functions of frequency and echo times:—

$$\theta_r = \omega t_r/2 \text{ radians} \quad .. \quad .. \quad .. \quad (1)$$

The individual echo networks are of two types, depending only on far-end terminations. The two-terminal impedances z_{11} and z_{12} are related, in the example, to the cases for open-circuited and short-circuited terminations respectively. The former produces a contribution of positive echo distortion and the latter a contribution of opposite sign. It can be shown quite readily that z_{11} and z_{12}, which are the impedances of the echo lines shunted by their respective characteristic resistances R_{11} and R_{12}, are given by:—

$$z_{11} = \frac{-jR_{11}\cot\theta_1}{1 - j\cot\theta_1} = \frac{R_{11}}{2}\left[(1 + \cos 2\theta_1) - j\sin 2\theta_1\right] \quad .. \quad (2)$$

$$z_{12} = \frac{jR_{12}\tan\theta_2}{1 + j\tan\theta_2} = \frac{R_{12}}{2}\left[(1 - \cos 2\theta_2) + j\sin 2\theta_2\right] \quad .. \quad (3)$$

which, in terms of t, are:—

$$z_{11} = \frac{R_{11}}{2}\left[(1 + \cos\omega t_1) - j\sin\omega t_1\right] \quad .. \quad (4)$$

$$z_{12} = \frac{R_{12}}{2}\left[(1 - \cos\omega t_2) + j\sin\omega t_2\right] \quad .. \quad (5)$$

Now the simulating impedance Z_s is simply:—

$$Z_s = R_o + z_{11} + z_{12} + z_{13} + \ldots \; z_{1n} \quad .. \quad (6)$$

In the case of the network of Fig. 1 this impedance can be expanded as follows:—

$$Z_s = \left[R_o + \frac{R_{11}}{2}(1 + \cos\omega t_1) + \frac{R_{12}}{2}(1 - \cos\omega t_2) + \frac{R_{13}}{2}(1 - \cos\omega t_3) \ldots + \frac{R_{1n}}{2}(1 + \cos\omega t_n)\right] - j\left[\frac{R_{11}}{2}\sin\omega t_1 - \frac{R_{12}}{2}\sin\omega t_2 - \frac{R_{13}}{2}\sin\omega t_3 \ldots + \frac{R_{1n}}{2}\sin\omega t_n\right] \quad .. \quad .. \quad .. \quad (7)$$

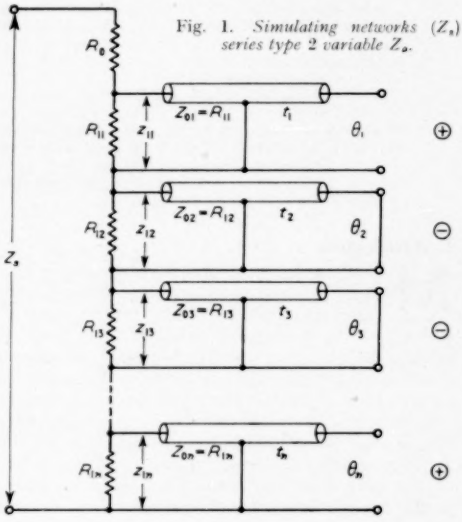

Fig. 1. *Simulating networks (Z_s) series type 2 variable Z_o.*

It has been shown in the previous paper[1] that the resistances R_{11} R_{12} $R_{13} \ldots R_{1n}$ can be expressed directly in terms of R_o and the coefficients k_1 k_2 $k_3 \ldots k_n$ which define the amplitudes of the separate echo components with respect to the amplitude of the undistorted signal. It should be noted here that the k coefficients are not intended to convey the signs of the echoes.

Any coefficient k_p defines a particular resistance R_{1p} by:—

$$R_{1p} = \frac{2k_pR_o}{1 - (k_1 + k_2 + k_3 + \ldots + k_p \ldots + k_n)} \quad .. \quad .. \quad .. \quad (8)$$

which will be written as:—

$$R_{1p} = \frac{2k_pR_o}{1 - \sum k} \quad .. \quad .. \quad .. \quad (9)$$

The complete expansion for Z_s in Fig. 1 is

then:—

$$Z_s = R_o + \frac{k_1 R_o(1 + \cos \omega t_1)}{1 - \sum k} + \frac{k_2 R_o(1 - \cos \omega t_2)}{1 - \sum k} \ldots + \frac{k_n R_o(1 + \cos \omega t_n)}{1 - \sum k} - j\left[\frac{k_1 R_o \sin \omega t_1}{1 - \sum k} - \frac{k_2 R_o \sin \omega t_2}{1 - \sum k} \ldots + \frac{k_n R_o \sin \omega t_n}{1 - \sum k}\right] \quad \ldots \quad (10)$$

which, in the general form, reduces to:—

$$Z_s = \frac{R_o\left(1 + \sum \pm k \cos \omega t\right)}{1 - \sum k} + \frac{jR_o \sum \mp k \sin \omega t}{1 - \sum k} \quad \ldots \quad (11)$$

Where alternative signs are used the upper is required for all individual terms resulting from positive echo components and the lower is required for negative components.

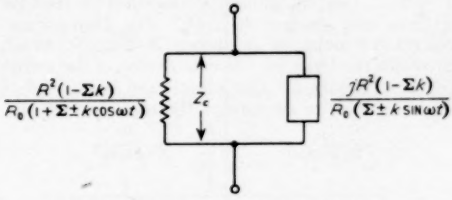

Fig. 2. *Basic correcting network Z_c.*

3. Transfer Constants

3.1. *Actual System or Simulating Network*

The amplitude transfer constant of the actual system is obtainable directly by taking the modulus of Z_s in equation (11).

$$|Z_s| = \frac{R_o}{1 - \sum k}\left[\left(1 + \sum \pm k \cos \omega t\right)^2 + \left(\sum \mp k \sin \omega t\right)^2\right]^{\frac{1}{2}} \quad \ldots \quad (12)$$

The quotient $R_o/\left[1 - \sum k\right)$ is resistive and constant and hence the system amplitude characteristic is given exactly by:—

$$\alpha_s = 10 \log_{10}\left[\left(1 + \sum \pm k \cos \omega t\right)^2 + \left(\sum \mp k \sin \omega t\right)^2\right] \text{ decibels.} \quad \ldots \quad (13)$$

relative to the system without echo distortions.

The system phase constant is simply:—

$$\beta_s = \tan^{-1}\left(\frac{\sum \pm k \sin \omega t}{1 + \sum \pm k \cos \omega t}\right) \quad \ldots \quad (14)$$

using the convention for transmission systems that a phase advance is considered as a negative phase shift.

The group delay of the system can be obtained from equation (14) by differentiation and has the value:—

$$\frac{d\beta_s}{d\omega} = \frac{\sin \beta_s}{\sum \pm k \sin \omega t}\left[\cos \beta_s\left(\sum \pm kt \cos \omega t\right) + \sin \beta_s\left(\sum \pm kt \sin \omega t\right)\right] \quad \ldots \quad (15)$$

Equation (11) can be written as:—

$$Z_s = \frac{R_o}{1 - \sum k}\left(1 + \sum \pm k e^{-j\omega t}\right) \quad \ldots \quad (16)$$

and equation (13) as:—

$$\alpha_s = 20 \log_{10}\left|\left(1 + \sum \pm k e^{-j\omega t}\right)\right| \text{ decibels} \quad (17)$$

It is important at this stage to consider the significance of the term $ke^{-j\omega t}$. This, in the most general case, will be called the E function which can be treated equally well as a function of frequency or of time.

It is now possible to define two forms of the E function which, taken separately, yield complete expressions for either the steady-state or the time characteristics of the system. The first expression uses the interpretation:—

$$E_f = ke^{-j\omega t} \quad \ldots \quad (18)$$

in which E_f is a frequency function which must be identified with the various and separate values of k and t. It is then possible to write equation (16) as:—

$$Z_s = \frac{R_o}{1 - \sum k}\left(1 + \sum \pm E_f\right) \quad \ldots \quad (19)$$

It is important now to realize that an associated E function can be defined which yields an equation almost identical with (19) but in which the interpretation is entirely in terms of time phenomena. If we define E_t as an operator of such a kind that the undistorted input current i_o can be transformed into an echo component of relative amplitude k displaced by a time t from the initial pulse, then $E_t i_o$ is a simple displaced echo component. We can now express the distorted output voltage as:—

$$V_d = \frac{R_o i_o}{1 - \sum k}\left(1 + \sum \pm E_t\right) \quad \ldots \quad (20)$$

which takes the form of an initial pulse followed by the appropriate train of echo components. In the following procedure it will be indicated that

an expression of the form of equation (20) can be treated in a simple algebraic fashion to yield a number of important results in complex systems.

3.2. *Correcting Network* Z_c

It has been stated previously that Z_c is the inverse of Z_s with respect to any arbitrarily chosen resistance R. The resistance and reactive components of Z_c can be found directly from equation (11) and are shown in the basic form in Fig. 2. The constant and time (or frequency) variable parts are separated in Fig. 3. The

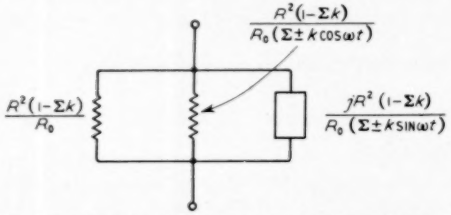

Fig. 3. *Separation of Z_c into constant and variable parts.*

insertion conditions for the network are shown in Fig. 4, in which the resistance r replaces the term $R^2(1 - \sum k)/R_o$, the constant resistance has been divided between source and load and the distorted current i_d is replaced by the equivalent e.m.f. $2\,ri_d$. This is the time equalizer connected under the practical conditions of finite source and load resistances such as would exist in a coaxial-cable system. It will be seen that if the actual system (or Z_s) presented no distortion, then all the coefficients $k_1\ k_2\ \ldots k_n$ would be zero and the whole of the network within the dotted line would be of infinite impedance at all frequencies and would have no effect on the transmission. This is the datum against which the Z_c attenuation must be assessed.

The complex voltage factor is $V_o/i_d r$ and a simple calculation shows that the value of this ratio is:—

$$\frac{V_o}{i_d r} = \frac{1}{1 + \sum \pm k \cos \omega t - j \sum \pm k \sin \omega t} = \frac{1}{1 + \sum \pm k e^{-j\omega t}} \quad .. \quad .. \quad .. \quad (21)$$

The value for the factor must be the reciprocal of the quantity defining the variation of Z_s with frequency [equations (11) and (16)]. The loss due to the time equalizer, as a function of frequency is, therefore, given by equation (17). The phase angle β_c is given by:—

$$\beta_c = -\beta_s \quad .. \quad .. \quad .. \quad (22)$$

and the group delay by:—

$$\frac{d\beta_c}{d\omega} = -\frac{d\beta_s}{d\omega} \quad .. \quad .. \quad .. \quad (23)$$

So far, in considering the insertion characteristics of the time-equalizer network Z_c, we have considered the attenuation as a function of frequency, but a somewhat unexpected result appears if we consider the attenuation of the waveform of the desired signal. In practical cases it is this reduction of the desired signal which is of interest, but the loss is not directly apparent from equation (17).

It is useful to maintain a clear interpretation of the physical process of pulse transmission through the Z_c system of Fig. 4. The actual form of the circuit in Fig. 4 can be illustrated conveniently by using the appropriate inversion of the simulating impedance Z_s of Fig. 1 and this is shown in Fig. 5. If the duration of the initial pulse or desired signal is short in time compared with the echo times of any of the delay components then it is easy to see that the voltage conditions across the complete Z_c network would be exactly the same if all the delay components were replaced by fixed resistances having values equal to the respective characteristic resistances R_{11}, R_{12}, R_{13} $\ldots$ R_{1n}. This transmission condition persists for a time not greater than t_1. The 'temporary' equivalent network is shown in Fig. 6, which provides the basis for the calculation of the initial pulse attenuation. The source and load form of Fig. 4 is used to emphasize the insertion effect.

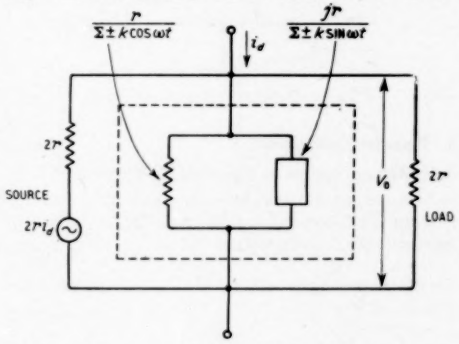

Fig. 4. *The insertion form of Z_c.*

The shunts in parallel with R^2/R_o are:—

$$\frac{2R^2}{R_{11}} = \frac{R^2}{R_o}\left(\frac{1-\sum k}{k_1}\right)$$
$$\frac{2R^2}{R_{12}} = \frac{R^2}{R_o}\left(\frac{1-\sum k}{k_2}\right)$$
$$\frac{2R^2}{R_{1n}} = \frac{R^2}{R_o}\left(\frac{1-\sum k}{k_n}\right) \quad .. \quad .. \quad .. \quad (24)$$

The parallel combination R_n of these shunts is:—

$$R_n = \frac{R_2}{R_o}\left(\frac{1-\sum k}{\sum k}\right) \quad .. \quad .. \quad .. \quad (25)$$

The voltage factor produced by these shunts is given by:—

$$\frac{\text{Output voltage (no correction)}}{\text{Output voltage (full correction)}} = \frac{1}{1-\sum k} \quad (26)$$

$$\text{or} \quad \alpha_c = 20 \log_{10}\left(\frac{1}{1-\sum k}\right) \text{ decibels.} \quad .. \quad (27)$$

This result should be compared with equation (17).

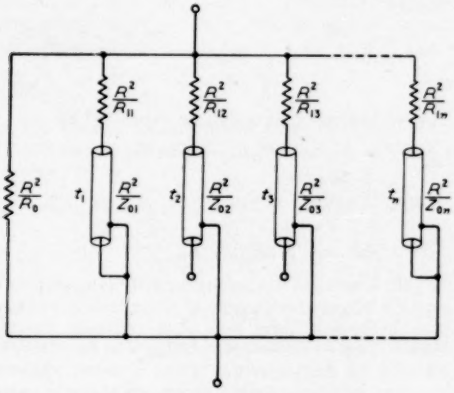

Fig. 5. *Z_c as the inverse of Z_s in Fig. 1.*

The desired signal is, therefore, attenuated by an amount which is not simply related to the attenuation undulations produced by Z_c on the amplitude/frequency characteristic. This finite reduction of the desired signal $(\sum k < 1{\cdot}0)$, as viewed on the time scale, is followed by infinite attenuation of the separate distortion components which follow later. This conception of a variable attenuation with time is not in any way related to non-linearity or strobe effects and it must not be considered as giving any more difficulty in concept than the more familiar points of infinite attenuation in amplitude/frequency characteristics.

4. Tandem System Characteristics

4.1. *General*

The properties of defined quadripoles connected in tandem, or of a sub-multiple of a defined quadripole, are easy to determine from steady-state characteristics (amplitude/frequency and phase/frequency) owing to the additive nature of such parameters. When the initial information has been given in the form of time characteristics the problems of manipulation have presented such difficulties that solutions, where attempted, have remained in the purely symbolic form, but some recent developments along interesting lines have been made by Raymond[2] and Gloess[3]. The technique which follows in this paper is related to the Gloess procedure, but one restriction in this has now been removed and the more general solution is developed.

In a previous paper[4] the author outlined the forms of the more important outstanding problems and these can be expressed as:—

(a) "Given the time characteristics of a linear quadripole, what are the time characteristics of a chain of m quadripoles of the same type connected in tandem?"

This problem is important in laboratory work on systems as the fraction of an actual system available for measurements is often very limited.

(b) "Given the time characteristics of a chain of m linear quadripoles connected in tandem, what are the time characteristics of one of these quadripoles considered alone?"

This problem appears when the performance specification of a complete system is given and it is necessary to fix limits for the construction and testing of one part of the system.

(c) Problem (c) is a modification of (a) to the extent that the m quadripoles need not be considered as alike.

This case is mainly of interest when a few known but unlike quadripoles are connected in tandem (e.g., radio link connected to coaxial-cable system).

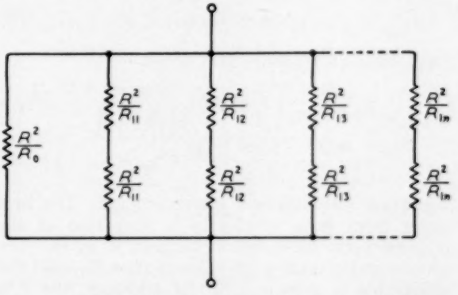

Fig. 6. *The 'temporary' equivalent of Z_c in Fig. 5.*

4.2. *Similar Quadripoles in Tandem [Problem (a)]*

The simplest case in this category appears when the quadripole distortion can be considered as a single echo. This is quite a practical example and

may be found in those systems where some aspect of the structure emphasizes an echo distortion with a particular delay. A coaxial-cable system could give this form of distortion if irregularities exist at the cable joints, which are usually at constant spacings.

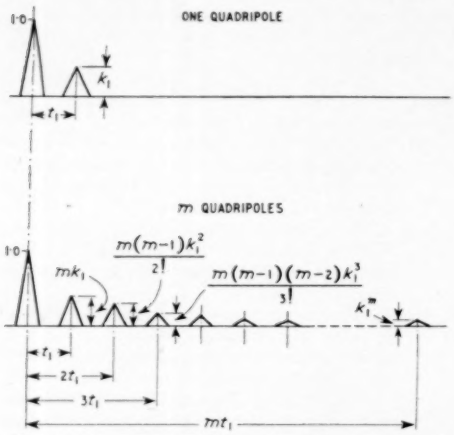

Fig. 7. *The effect of transmission through m simple quadripoles.*

In terms of the simulating network the single quadripole can be represented as a form of equation (20), viz:

$$V_d = \frac{R_o i_o}{1 - k_1}\left(1 + E_{t_1}\right) \quad .. \quad .. \quad (28)$$

and the system of m quadripoles as:—

$$\left(V_d\right)_m = \frac{R_o i_o}{1 - k_1}\left(1 + E_{t_1}\right)^m \quad .. \quad .. \quad (29)$$

Expansion of equation (29) gives:—

$$\left(V_d\right)_m = \frac{R_o i_o}{1 - k_1}\left[1 + mE_{t_1} + \frac{m(m-1)}{2!}E^2{}_{t_1} + \frac{m(m-1)(m-2)}{3!}E^3{}_{t_1} + \ldots.\right] \quad (30)$$

Equation (30) requires interpretation. The first order term mE_{t_1} is clearly a distortion of the original form increased in amplitude by m. The second-order term is an echo at time $2t_1$, and the amplitude is $m(m-1)/2$. In addition, the $E^2{}_{t_1}$ term includes a coefficient $k_1{}^2$. Fig. 7 shows the waveform of the system of m quadripoles in tandem. Even if m is large the combined coefficients are fairly rapidly convergent if $k_1 m$ is less than unity—which would certainly be the case with most practical systems. It is interesting to note the analogy with the mechanism of harmonic production in a non-linear system under the excitation of a single (sinusoidal) component.

The general case of m multi-echo quadripoles in tandem can be solved along the lines indicated above but the resultant waveform distortion is much more complex. The single quadripole is defined as:—

$$V_d = \frac{R_o i_o}{1 - \sum k}\left(1 + \sum E_t\right) \quad .. \quad .. \quad (31)$$

and the complete system as:—

$$\left(V_d\right)_m = \frac{R_o i_o}{1 - \sum k}\left(1 + \sum E_t\right)^m \quad .. \quad (32)$$

Now:—

$$\left(1 + \sum E_t\right)^m = 1 + m\sum E_t + \frac{m(m-1)\left(\sum E_t\right)^2}{2!} + \ldots. \quad .. \quad .. \quad .. \quad (33)$$

The second-order term can be expanded as:—

$$\begin{aligned}\left(\sum E_t\right)^2 &= E_{t_1}{}^2 + 2E_{t_1}E_{t_2} + 2E_{t_1}E_{t_3} + \ldots. \\ &\quad + 2E_{t_1}E_{tn} \\ &+ E_t{}^2{}_2 + 2E_{t_2}E_{t_3} + 2E_{t_2}E_{t_4} \ldots. + 2E_{t_2}E_{tn} \\ &\quad + \ldots.... \\ &\quad + E_t{}^2{}_{(n-1)} + 2E_{t(n-1)}E_{tn} \\ &\quad + E^2{}_{tn} \\ &= \left(\sum E_t{}^2 + 2\sum E_t E_t\right) \quad .. \quad .. \quad .. \quad (34)\end{aligned}$$

Similarly the third-order term can be written as:—

$$\left(\sum E_t\right)^3 = \left(\sum E_t{}^3 + 3\sum E_t{}^2 E_t + 6\sum E_t E_t E_t\right) \quad (35)$$

As each one of the E_t functions contains a k coefficient it will be apparent that terms higher than the third order will be seldom of any practical importance. Substituting in equation (33) the values of $\left(\sum E_t\right)^2$ and $\left(\sum E_t\right)^3$ from equations (34) and (35) respectively, we obtain:—

$$\begin{aligned}\left(1 + \sum E_t\right)^m &= 1 + m\left(\sum E_t\right) + \frac{m(m-1)}{2!}\left[\sum E_t{}^2 + 2\sum E_t E_t\right] \\ &+ \frac{m(m-1)(m-2)}{3!}\left[\sum E_t{}^3 + 3\sum E_t{}^2 E_t + 6\sum E_t E_t E_t\right] + \ldots.. \quad .. \quad .. \quad (36)\end{aligned}$$

in which $\sum E_t E_t$ represents the combination of all terms such as $E_{tp}E_{tq}$, $\sum E_t{}^2 E_t$ the combination of all terms such as $E_{tp}{}^2 E_{tq}$ and so on. A particular term of the form $E_{tp}{}^2 E_{tq}$ would be associated with a complete coefficient,

$$\frac{m(m-1)(2m-1)}{2}k^2{}_p k_q$$

and the distortion of this amplitude would be located at a time $(2t_p + t_q)$ after the desired signal.

Equation (36) gives the complete and convergent solution of Problem (a). Again the analogy will be seen with the non-linearity problem in which such terms as $E_{tp}E_{tq}$ would represent intermodulation terms of the second order.

4.3. *Sub-Multiple of a System [Problem (b)]*

The characteristics of a fraction of a complete system are obtained by the substitution of $\frac{1}{m}$ for m in the above cases for multiples. For example, one mth of a complete system characterized by a single echo distortion would be defined by:—

$$\left(V_d\right)_{\frac{1}{m}} = \frac{R_o i_o}{1 - k_1}\left(1 + E_{t_1}\right)^{\frac{1}{m}} \quad .. \quad (37)$$

which should be compared with equation (32). Expansion of equation (37) gives:—

$$\left(V_d\right)_{\frac{1}{m}} = \frac{R_o i_o}{1 - k_1}\left[1 + \frac{E_{t_1}}{m} - \frac{(m-1)}{2!m^2}E_{t_1}^2 + \frac{(m-1)(2m-1)}{3!m^3}E_{t_1}^3 \ldots\right] \quad .. \quad (38)$$

The physical interpretation of equation (38) is curious at first sight, as it means that a complete system showing only one echo distortion can comprise m like quadripoles each of which would produce, from a single initial signal, an infinite array of echo distortions. Here again the analogy with a non-linear system is interesting as it is known that if the output of such a system (defined by an infinite polynomial) is a single component (i.e., sinusoidal) then the input must be represented by an infinite series comprising a fundamental and all orders of harmonics.

Fig. 8, which should be compared with Fig. 7, shows the waveform of one mth of a complete system with one echo.

When the system is characterized by a complex array of echo distortions then the mth part is defined by:—

$$\left(V_d\right)_{\frac{1}{m}} = \frac{R_o i_o}{1 - \sum k}\left(1 + \sum E_t\right)^{\frac{1}{m}} \quad .. \quad (39)$$

The bracket can be expanded to give:—

$$\left(1 + \sum E_t\right)^{\frac{1}{m}} = 1 + \frac{1}{m}\left(\sum E_t\right) - \frac{(m-1)}{2!m^2}\left[\sum E_t^2 + 2\sum E_t E_t\right] + \frac{(m-1)(2m-1)}{3!m^3}\left[\sum E_t^3 + 3\sum E_t^2 E_t + 6\sum E_t E_t E_t\right] \ldots\ldots \quad .. \quad .. \quad (40)$$

which is a complete solution of Problem (b). The definitions are the same as those applied to equation (36).

4.4. *Tandem Connection of Unlike Quadripoles [Problem (c)]*

This problem is easier than might be expected. If the individual quadripoles are characterized by echo distortions arising from $\sum E_t'$, $\sum E_t''$, $\sum E_t'''$, etc., then the overall system can be defined by:—

$$\left(V_d\right)_{\text{TOTAL}} = \frac{R_o i_o}{f(k)}\left(1 + \sum E_t'\right)\left(1 + \sum E_t''\right)\left(1 + \sum E_t'''\right) \ldots\ldots \quad .. \quad .. \quad (41)$$

$$= \frac{R_o i_o}{f(k)}\left[\left(1 + \sum E_t' + \sum E_t'' + \sum E_t''' + \ldots.\right) + \left(\sum E_t'\right)\left(\sum E_t''\right) + \left(\sum E_t'\right)\left(\sum E_t'''\right) + \ldots. + \left(\sum E_t'\right)\left(\sum E_t''\right)\left(\sum E_t'''\right) + \ldots. + \ldots\ldots\ldots\right] \quad .. \quad .. \quad .. \quad (42)$$

Products such as $\left(\sum E_t'\right)\left(\sum E_t''\right)$ can be expanded by the methods indicated in the section above.

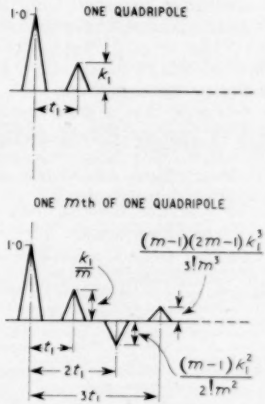

Fig. 8. *The effect of transmission through one* m*th of a system.*

5. General Considerations

It will have been noted in Section (4) that the first-order distortion in a system of m quadripoles is the sum of the distortions produced by the parts when subjected individually to the same signal as that applied to the complete system. This additive characteristic can be considered as

accurate enough when the echo distortions are small. Departure from this approximation can be assessed by the inclusion of the second- and higher-order terms in the convergent series given in the various cases.

It is sometimes thought that the waveform distortions produced by some systems are not of a kind suitable for correction by 'Time Equalizers'. This limitation can be true, but it means little more than that amplitude/frequency and phase/frequency correction should first be applied before 'Time Equalizers' can be used in their proper role. For example, it would be pointless, or very uneconomic technically, to attempt a correction of the smooth $(\omega)^{\frac{1}{2}}$ attenuation characteristic of a coaxial cable by waveform networks. It is required that there should be a value set for the maximum frequency f_m to be transmitted, and a smooth correction (up to this frequency) applied by the use of conventional steady-state networks. The residual undulations of the characteristic should give differences from a flat characteristic, up to f_m, which sum approximately to zero in this range. If this criterion is met then the residual distortions can be represented by an array of echos displaced from the unit pulse of minimum width which can be expected to pass through the system. This 'best' pulse would be approximately of raised-cosine form of width equal to $1/2f_m$ at half height.

Any echo distortion displaced from the 'best' pulse by less than about $1/2f_m$ would show as a droop across the characteristic up to f_m—and this effect, in a limited-bandwidth system, should be corrected by steady-state methods.

It is very desirable that a close watch should be maintained for any mechanism in the system by which distortion of an echo character can be produced. At least, these distortions should be corrected by waveform methods.

REFERENCES

[1] D. C. Espley, "The Exact Solution for Transient Distortion in Networks," *Electronic Engineering*, March 1950, Vol. 22, No. 265, pp. 82-87.

[2] F. H. Raymond, "Régimes Transitoire et Technique des Impulsions (1)", *L'Onde Electrique*, June 1948, Vol. 28, No. 255, pp. 222-234.

[3] P. F. Gloess, "Suggestions du Groupe d'Etude des Régimes Transitoire de le Société des Radioelectriciens," Paris, May 1947.

[4] D. C. Espley, "Système de Télécommunications en Régime Transitoire," *L'Onde Electrique*, Vol. 27, No. 249, pp. 437-446. (Presented before the Société Française des Electriciens, Paris, May 1947).

M.K.S. UNITS

At a meeting in Paris during July 1950, the Technical Committee No. 24 (Electrical and Magnetic Units) of the International Electrotechnical Commission made certain decisions about the m.k.s. system of units. They are:—

1. That the Newton is finally adopted as the unit of force.
2. That it is convenient to base the system for electrical quantities on four principal units and that it is recommended that, for the purpose of developing the definitions of the units, the fourth principal unit should be the ampere.
3. That rationalization be used, which leads to:—

$\mu_o = 4\pi \times 10^{-7}$ H/m for the permeability of free space and,

$\epsilon_o = 1/\mu_o c^2$ F/m for the permittivity of free space, where c is the velocity of light in free space (ϵ is the I.E.C. symbol for permittivity).

These decisions are recorded in the official minutes of the meeting and an extract CN(ELE) 981 has been prepared by the British Standards Institution which is the British National Committee of the I.E.C.

CORRESPONDENCE

Star-Delta Theorem

SIR,—As one who has never found it easy to remember the formulation of the $Y\ \nabla$ transformation, I was pleased to find that it can be expressed diagrammatically as follows. Fig. 1 shows the mesh equivalent of a star which

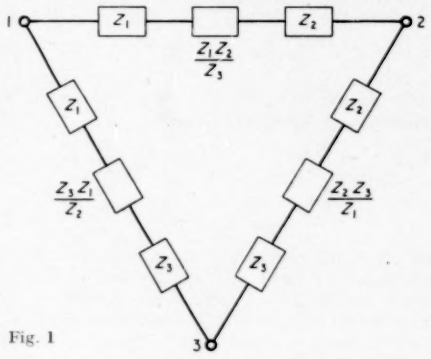

Fig. 1

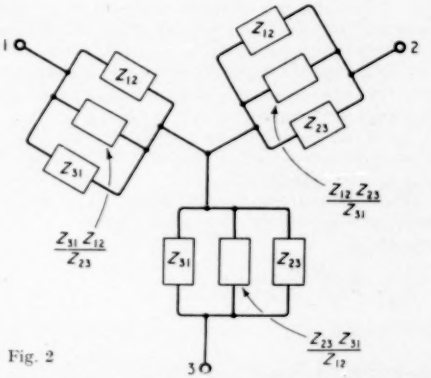

Fig. 2

has terminals 1, 2 and 3 with impedances Z_1, Z_2 and Z_3 connected from these terminals to a common point. Fig. 2 shows the star equivalent of a mesh which has terminals 1, 2 and 3 with impedances Z_{12}, Z_{23}, Z_{31}, connected between these terminals.

EMRYS WILLIAMS.

University College of North Wales,
Bangor,
15th June, 1951.

ABSTRACTS and REFERENCES

Compiled by the Radio Organization of the Department of Scientific and Industrial Research, and published by arrangement with that Department.

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

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ACOUSTICS AND AUDIO FREQUENCIES

534 **1811**

Program of the Fortieth Meeting of the Acoustical Society of America (Boston, Mass., 9th–11th November, 1950).—(*J. acoust. Soc. Amer.*, Jan. 1951, Vol. 23, No. 1, pp. 142–150.) Summaries are given of the papers presented, which included the following:—

A5. Attenuation of Sound in Water containing Air Bubbles.—D. T. Laird & P. M. Kendig.

B5. On the Relation between the Sound Fields Radiated and Diffracted by Plane Obstacles.—F. M. Wiener.

B6. Synthesis of Line Source Directivity Patterns.—R. Hills, Jr.

B7. Viscosity Effects in Acoustic Inductances.—A. W. Nolle.

B8. The Non-Linear Interaction of a Plane Progressive Wave with a Small Sphere.—P. Westervelt.

B9. First- and Second-Order Acoustic Fields with Viscosity and Relaxation.—J. J. Markham.

C3. Comparison of Absorption Coefficients Measured in a Reverberation Chamber with Statistical Coefficients from Impedance Values.—H. C. Hardy, F. G. Tyzzer, L. G. Ramer & J. E. Ancell.

C4. Damping Effects of Porous Absorbing Units in Rooms.—R. W. Roop.

C5. The Record-Rerecord Test for Rooms.—W. Rudmose.

D2. The Transmission of Sound through Single and Double Walls at Oblique Incidence.—R. E. Beatty, Jr, R. H. Bolt & J. Young.

D3. On the Design of Sound Absorbing Ducts.—U. Ingård, J. J. Baruch, L. L. Beranek & S. Labate.

D4. A Wide-Range Acoustic Termination.—O. K. Mawardi.

D5. On the Non-Specular Reflection of Sound from Absorbent Surfaces.—V. Twersky.

D6. A Comparison of Sound Pressures Developed in Earphone Couplers and in the Ear.—E. L. R. Corliss, R. F. Brown & K. T. Lemmon.

F1. New Assembly for the Measurement and Analysis of Sound.—L. W. Sepmeyer.

F2. Non-Linear Distortion in Hearing Aids.—A. Peterson.

F3. An Automatic Non-Linear Distortion Analyzer.—H. F. Olson & D. F. Pennie.

F4. A Photographic Method for Displaying Sound Wave Space Patterns.—W. E. Kock & F. K. Harvey.

F5. A Feedback-Controlled Calibrator for Phonograph Pick-Ups.—J. G. Woodward.

F6. Equipment for the Field Measurement of Reverberation Times.—W. Rudmose & H. Mims.

F7. Higher Order Gradient Microphones using a Single Diaphragm.—M. E. Hawley & A. H. Kettler.

534 : 016 **1812**

References to Contemporary Papers on Acoustics.—A. Taber Jones. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 240–247.) Continuation of 1541 of July.

534 : 577 **1813**

A Bibliography on Sonic and Ultrasonic Vibration: Biological, Biochemical and Biophysical Applications.—G. M. Naimark, J. Klair & W. A. Mosher. (*J. Franklin Inst.*, Feb. & March 1951, Vol. 251, Nos. 2 & 3, pp. 279–299 & 402–408.) A list of 580 references for the period 1900–1950, grouped according to the year of publication and with both author and subject index.

534.22-14 **1814**

The Velocity of Sound in Sea Water.—A. Weissler & V. A. Del Grosso. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 219–223.) Full paper. Summary abstracted in 514 of March.

534.231 **1815**

A New Expansion for the Velocity Potential of a Piston Source.—A. H. Carter & A. O. Williams, Jr. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 179–184.) Full paper. Summary noted in 521 of March.

534.231 **1816**

On the Acoustical Radiation of an Emitter Vibrating in an Infinite Wall.—J. Pachner. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 185–198.) The velocity-potential distribution for a circular emitter in an infinite wall is calculated by the King method for points immediately in front of the wall.

534.231 **1817**

On the Acoustical Radiation of an Emitter Vibrating Freely or in a Wall of Finite Dimensions.—J. Pachner.

(*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 198–208.) The radiation field of a source vibrating freely or in a finite wall is considered as made up of the field of the same source in an infinite wall together with a field that can be computed from an integro-differential equation that follows from Rayleigh's formula. An abstract solution of the problem is given which can be applied for any shape of wall or wavelength of sound. Computation will be easier if the wall is circular and if the wavelength is large in comparison with the wall dimensions.

534.232 : 621.315.612.4 **1818**

A Barium Titanate Transducer Capable of Large Motion at an Ultrasonic Frequency.—W. P. Mason & R. F. Wick. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 209–214.) Full paper. Summary abstracted in 560 of March, for which the above U.D.C. number is preferable.

534.241 **1819**

Analysis of Multiple-Echo Effect Arising from the Release of a Stored Wave Train.—L. Gold. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 214–218.) Calculation of the dependence of the number of observable echoes on the threshold sensitivity of the detector and on the absorption and reflection coefficients of the storage medium.

534.321.7 **1820**

Standard Musical Pitch.—F. W. Alexander. (*B.B.C. Quart.*, Spring 1951, Vol. 6, No. 1, pp. 62–64.) The history of standard pitch is sketched briefly, and the method used by the B.B.C. for deriving from a crystal-controlled 1-Mc/s oscillator the 440-c/s international-standard-pitch one broadcast before the start of the Third Programme is described.

534.321.9 : 534.231-14 **1821**

Measurements of the Underwater Sound Field Generated by Quartz Transducers.—W. Keck, G. S. Heller & A. O. Williams, Jr. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 168–172.) The pressure distributions in the ultrasonic fields produced in water by quartz transducers of various shapes were displayed on a c.r.o. and photographed. The ultrasonic beams, pulsed at a.f., were swept backward and forward across a small microphone placed at various distances from the source. The microphone output was rectified and amplified before being applied to the c.r.o. All the transducers were of X-cut quartz about 2 cm in diameter, excited near resonance at about 1 Mc/s. The results obtained show that simple piston theory is applicable, provided the baffles and electrodes used satisfy certain geometrical conditions.

534.321.9 : 534.232 **1822**

High-Power Ultrasonic Siren.—L. Pimonow. (*Ann. Télécommun.*, Jan. 1951, Vol. 6, No. 1, pp. 23–26.) The siren described was developed at the C.N.E.T., and is of the type in which a toothed rotor interrupts a powerful air current emerging from holes in a stator [see also 917 of 1948 (Allen & Rudnick)]. The frequency range is 5–25 kc/s, and a radiated acoustic power > 2 kW can be obtained by introducing compressed nitrogen into the casing. Applications are discussed.

534.321.9 : 534.614-13 **1823**

Ultrasonic Velocities in Gases at Low Pressures.—R. A. Boyer. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 176–178.) Measurements of velocity as a function of pressures down to about 2 mm Hg were made at a frequency of 970 kc/s. At the lowest pressure the following increases in velocity over the values under standard conditions were observed: A, 27%; N_2, 16%; O_2, 20%; dry CO_2-free air, 7%.

534.321.9 : 621.395.61 **1824**

Effects of Reflected Signals and Electric Pick-Up at an Ultrasonic Microphone.—A. O. Williams, Jr, & W. Keck. (*J. acoust. Soc. Amer.*, March 1951, Vol. 23, No. 2, pp. 173–175.) The mixing process is analysed for a particular arrangement of equipment, and experimental evidence is adduced in support of the conclusions that (*a*) the main response pattern of the microphone shows space variations of the order of both λ and $\lambda/2$, (*b*) the $\lambda/2$ variation dies out, at sufficiently great distances from the source, at a rate showing that a single echo path together with a constant electric signal is responsible, (*c*) the direct, reflected, and electric-pickup signals can be resolved analytically.

534.374 **1825**

⅓-Octave Variable Filter.—G. Fontanellaz. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, 1st Feb. 1951, Vol. 29, No. 2, pp. 48–51. In French and German.) This band-pass filter, designed for use in reverberation measurements, comprises a Zobel-type variable LC network. The range covered is 50 c/s–10 kc/s, using a 9-position and a 3-position switch respectively for octave and ⅓-octave selection. The circuit arrangement is described and attenuation characteristics are shown.

534.6 : 621.395.625 **1826**

Modulation Measurement at Points of Frequency-Dependent Maximum-Amplitude.—K. H. R. Weber. (*Frequenz*, Nov. 1950, Vol. 4, No. 11, pp. 295–298.) The use of frequency compensation in the meter circuit in order to obtain a true indication of amplitude is described, with reference to the response curves of optical and disk recorders.

534.78 **1827**

Representations of Speech Sounds and some of their Statistical Properties.—Sze-Hou Chang, G. E. Pihl & M. W. Essigmann. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 147–153.) 1950 I.R.E. National Convention paper noted in 1319 of 1950. Properties of speech sounds and methods of transformation and analysis such as infinite clipping, autocorrelation and spectrographic analysis are discussed with a view to the identification of essential elements of speech intelligibility.

534.846.3/.4 **1828**

Sound Reinforcing.—D. W. Pipe. (*Wireless World*, March 1951, Vol. 57, No. 3, pp. 117–119.) The sound-distribution system installed in the Assembly Hall and three adjacent halls at Church House, Westminster, is described. In the Assembly Hall low-level reinforcement from small loudspeakers secured to the seat backs is provided. The facilities installed include four-channel simultaneous interpretation; any member of the audience can select the interpretation he requires by operating a switch fitted in the loudspeaker panel, which controls the input to his headphones. Any speaker on the floor of the hall is within the zone of one of the eleven suspended microphones, the appropriate one being selected by the system operator in the balcony, who has an unobstructed view of the assembly.

534.846.3/.4 : 621.395.623.7 **1829**

From Decentralized to Central Sound Transmission.—F. Spandöck. (*Elektrotech. Z.*, 15th Feb. 1951, Vol. 72, No. 4, pp. 101–104.) The advantages of loudspeaker arrays as compared with distributed arrangements of loudspeakers are: lower installation costs; less amplifier power required; absence of zones of confusion; coincidence of directions of sight and sound. Microphone arrays have corresponding advantages, and when used in theatres need not be restrictively close to the actors. In sound-

reinforcement systems the sharp directivity of the arrays lessens the danger of acoustic feedback. A formula is given for the amplification at which feedback commences; the calculated values are compared with measurements. Stereophonic applications of arrays are briefly considered.

534.844.1 **1830**

Automatic Apparatus for Measurement of Reverberation Time.—F. J. v. Leeuwen. (*Tijdschr. ned. Radiogenoot.*, Jan. 1951, Vol. 16, No. 1, pp. 13–36. Discussion, p. 37.) The theory of reverberation time and its measurement is outlined, and apparatus for providing a moving-strip record of the reverberation-time/frequency curve of a room is described. By means of a relay arrangement a sound source within the room is periodically actuated and is interrupted when a steady state is reached. A microphone has its output circuit arranged so as to actuate a 'reverberation-time relay' when the sound level drops to 40 db below the level prevailing on interruption of the source. Actuation of this relay causes reversal of the motion of the recording pen of a decibel meter, the distance between turning point and zero axis giving the reverberation time. A frequency scale is provided on the record by means of a Wien-bridge circuit connected to the tone generator.

534.85 **1831**

Thorn Needles.—C. E. Watts : S. Kelly. (*Wireless World*, March 1951, Vol. 57, No. 3, pp. 121–122.) Comments on 806 of April (Pollock).

534.85/.86 : [621.396.645.029.3 + 621.395.623.7 **1832**

The FAS Audio System: Parts 3 & 4.—M. B. Sleeper. (*FM-TV*, Dec. 1950, Vol. 10, No. 12, pp. 24–26 & Feb. 1951, Vol. 11, No. 2, pp. 32–35, 38.) Description of (*a*) methods of adapting the FAS loudspeaker system to any good amplifier, (*b*) the construction of an 8-ft air-coupler, (*c*) experiments with 15-in. loudspeakers for high volume output. Parts 1 & 2: 805 & 804 of April.

534.86 : 621.396.712 **1833**

Developments in Studio Design.—L. L. Beranek. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1951, Vol. 12, No. 1, pp. 21–25.) Reprint. See 1847 of 1950.

621.3.018.78† : 621.395.613.3 **1834**

Graphical Representation of Nonlinear Distortion.—(*Radio tech. Dig., Édn franç.*, 1950, Vol. 4, No. 5, pp. 279–290.) French version of paper abstracted in 2964 of 1950 (Bressi & Sacerdote).

621.395.623.7 **1835**

Improving Loudspeaker Response with Motional Feedback.—R. L. Tanner. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 142–240.) A negative feedback voltage is obtained from a feedback coil wound over the existing voice coil in a conventional loudspeaker. This voltage, induced by the movement of the voice coil, is fed back to the driving amplifier. Mechanical and magnetic data for the design of an amplifier/loudspeaker combination are listed for five typical loudspeakers.

621.395.623.7 : 621.3.018.78† : 621.317.79 **1836**

An Automatic Nonlinear Distortion Analyzer [for loudspeakers].—H. F. Olson & D. F. Pennie. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 35–44.) The system consists of the standard automatic recorder for obtaining the frequency response curve, combined with a set of switched high-pass filters which suppress the fundamental and pass the harmonics. The overall-amplitude/frequency characteristic and the distortion-amplitude/frequency characteristic are recorded on the same sheet, and the percentage distortion at any frequency is obtained by comparing the two curves.

621.395.625.3 **1837**

Magnetic Recording with A.C. Bias.—R. E. Zenner. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 141–146.) The function of a.c. bias in magnetic recording is analysed in a manner similar to that used to explain a.m. Certain simplifying assumptions are made to facilitate manipulation of mathematical expressions. The analytical results are compared with experimental observations of harmonic distortion, amplitude of fundamental, spurious recorded frequencies, frequency response, difficulty of erasure, etc.

621.395.625.6 **1838**

Electron-Optical Sound-Recording Device.—P. G. Tager. (*C. R. Acad. Sci. U.R.S.S.*, 21st Aug. 1950, Vol. 73, No. 6, pp. 1181–1183. In Russian.) An arrangement of two c.r. tubes with x-deflections synchronized at a frequency much higher than the highest a.f. One tube has also a y-deflection system controlled by microphone currents, and has a modulation-controlling mask of some desired pattern interposed in front of a signal plate. Film is traversed in the y direction past the fluorescent screen of the second tube, whose beam is switched on or off according as current is or is not flowing in the signal-plate circuit, as controlled by the mask.

621.395.92 **1839**

A Master Hearing Aid.—E. Aspinall. (*J. Brit. Instn Radio Engrs*, Feb. 1951, Vol. 11, No. 2, pp. 45–50.) Description of a speech transmission system with widely variable characteristics, suitable for speech tests on deaf patients. Automatic-level-control, peak-limitation and attenuation-distortion circuits can be selected as desired, the merits of any transmission path being assessed by analysis of sound articulation scores.

621.395.92 **1840**

The Design of Commercial Hearing Aids.—J. P. Ashton. (*J. Brit. Instn Radio Engrs*, Feb. 1951, Vol. 11, No. 2, pp. 51–59.) A description of the engineering technique adopted in the design and construction of hearing aids with a view to providing instruments of the smallest possible size. Details of components and circuit elements are given.

AERIALS AND TRANSMISSION LINES

621.392.26† **1841**

Cut-Off Frequency in Two-Dielectric Layered Rectangular Waveguides.—J. van Bladel & T. J. Higgins. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 329–334.) Equations for the modes and eigenvalues are derived for a guide containing a single dielectric layer, a dielectric sandwich, or an air sandwich. Cut-off frequencies are shown graphically for a range of geometric and dielectric parameters.

621.392.26† : 621.396.52 **1842**

U.H.F. Filters and Lenses.—P. Marié. (*Radio tech. Dig., Édn franç.*, 1950, Vol. 4, Nos. 5 & 6, pp. 291–312 & 359–364, & 1951, Vol. 5, No. 2, pp. 97–109.) The analogy between optical and radio waves and the general theory of transparency are discussed, with particular reference to a system consisting of a series of metal diaphragms in a waveguide. Expressions derived for attenuation, etc., are applied to the problems of design of a band filter, losses being assumed negligible; a numerical example is given. Effects of losses due to resistance, and of multiple reflections in the transient regime, are analysed. The theory of e.m. lenses is considered and the required phase rotation is determined. The elements of calculation are indicated for the design of lenses of the pseudo-dielectric, the accelerating and the retarding types.

621.392.26† : 621.396.67 **1843**

Radiation Characteristics of Open Rectangular Waveguides (Horns) for Fundamental and Harmonic Oscillations in the H-Mode.—W. Reichardt. (*Frequenz*, Jan. 1951, Vol. 5, No. 1, pp. 23–27.) Theoretical determination of the radiation field for the H_{mn} mode for different values of m and n.

621.396.67 **1844**

The Field Radiated by Circular and Square Helical Beam Antennas.—H. L. Knudsen. (*Trans. Dan. Acad. tech. Sci.*, 1950, No. 8, 55 pp.) Theoretical investigations as well as measurements have shown that the current distribution on a helical aerial radiating in the axial mode may be described, for the greater part of the helix, as a progressive current wave with constant amplitude and phase velocity moving along the wire away from the feed point. Formulae for the e.m. field radiated from an arbitrary current distribution are reviewed and applied to the calculation of the field radiated by a circular helical aerial. The field is expressed in the form of rapidly converging series. The general formulae are used to derive the field of a helix with an integral number of turns, and a numerical example is worked out. Corresponding calculations are made for a square helical aerial, the field components being obtained as closed expressions from which approximate formulae for the circular helical aerial are derived. The series obtained for the circular helix are, however, so rapidly convergent and the closed expressions for the square helix are so simple that approximate formulae are hardly necessary.

621.396.67 : 621.397.6 **1845**

TV Receiving Antenna Research, Design and Production.—R. G. Peters. (*TV Engng, N.Y.*, Feb. 1951, Vol. 2, No. 2, pp. 8–10.) A report of methods used by some U.S. manufacturers.

621.396.671.4 **1846**

Complex Radiation Impedance of an Aerial System Electromagnetically Coupled to Another Aerial System.—R. G. Mirimanov. (*C. R. Acad. Sci. U.R.S.S.*, 21st Aug. 1950, Vol. 73, No. 6, pp. 1177–1179. In Russian.) Formulae are derived for the case of two systems each comprising a linear oscillator and a reflector having the form of a surface of revolution.

621.396.677 **1847**

Gain of Electromagnetic Horns.—W. C. Jakes, Jr. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 160–162.) An experimental investigation is described. For the horns tested it was found (*a*) that edge effects are $< 0{\cdot}2$ db, so that the gain may be computed to within that accuracy from their physical dimensions, using Schelkunoff's curves (2295 of 1943); (*b*) that for the transmission of power between two horns the ordinary transmission formula is valid, provided that the distance between the horns is measured between suitable reference points on the horns and not between their mouths.

621.396.677 : 538.56 : 535.312 **1848**

Experimental Determination of the Reflection Coefficient of Metal-Plate Media.—Ruze & Young. (See 1880.)

621.396.677 : 538.56 : 535.312 **1849**

Reflection and Transmission at the Surface of Metal-Plate Media.—Lengyel. (See 1879.)

621.396.677 : 621.396.9 **1850**

Microwave Radar Antenna.—H. N. Chait. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 103–105.) The polarization characteristics are determined in the transmission line system, where a magic-T junction is used with a variable phase shifter. Any arbitrarily chosen elliptical polarization can be transmitted and received by proper adjustment of the phase shift and rotation of the aerial feed. A circularly polarized search radar system can be used with a single aerial, as the transmitter for a circular component and the receiver for the reflected circular component are located at different terminals of the junction.

CIRCUITS AND CIRCUIT ELEMENTS

517.92 : 621.396.611.1 **1851**

Topological Methods for Investigating Problems relating to Nonlinear Oscillations.—T. Vogel. (*Ann. Télécommun.*, Jan. 1951, Vol. 6, No. 1, pp. 2–10.) The methods considered are useful for giving a physical insight into the phenomena corresponding to nonlinear differential equations, and for demonstrating the influence of various parameters. Work done by Poincaré, Bendixson and Dulac is surveyed, and the concepts of singular points, limit cycles, etc., are discussed. The method is used to investigate the solutions of nonlinear second-order differential equations having applications in physics, and is particularly suitable for dealing with discontinuous phenomena such as sawtooth oscillations. Topographical systems can be displayed on c.r.o. screens.

517.93 : 621.319.55 **1852**

On Amplitude Bounds for Certain Relaxation Oscillations.—N. Wax. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 278–281.) Amplitude limits are derived for the unique periodic solution of the generalized Lienard equation

$$\ddot{x} + f(x)\dot{x} + g(x) = 0$$

when the functions $f(x)$ and $g(x)$ are suitably restricted. The restrictions are less severe for the lower than for the upper limits. A class of equations is noted which includes the van der Pol equation and satisfies all the restrictions. Numerical results are given for the van der Pol equation.

621.314.2 + 621.318.42].003.1 **1853**

Minimum-Cost Transformers and Chokes.—H. C. Hamaker & T. Hehenkamp. (*Philips Res. Rep.*, Oct. 1950, Vol. 5, No. 5, pp. 357–394.) The theoretical design of a transformer is governed by the dependence of the price on the dimensions, and the way the apparent power and losses depend on the dimensions, magnetic-flux density and electric-current density. The problem of minimum cost, when power and losses are prescribed, is considered for various types of transformer. The solution is given in tabular form for practical application, and examples are calculated. This solution is, however, unsatisfactory in that it gives a design and price dependent only on the ratio of power to losses and not on their absolute magnitudes. Modifications of the theory to take account of limitations imposed by maximum temperature or flux-density are discussed.

621.318.4 : 621-05 **1854**

Universal Coil Winding.—E. Watkinson. (*J. Brit. Instn Radio Engrs*, Feb. 1951, Vol. 11, No. 2, pp. 61–69.) Reprint. See 2993 of 1950.

621.392.012.8 **1855**

Study of a Meshed Hydraulic Network, by the Method of Equalization of Loads: Reduction to an Equivalent Electrical Network.—C. Dubin & M. Magnien. (*Rev. gén. Élect.*, Feb. 1951, Vol. 60, No. 2, pp. 70–74.)

621.392.5 : 621.3.015.3 : 518.3 **1856**

Spectrum Analysis of Transient-Response Curves.—H. A. Samulon. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 175–186.) A method of computing the amplitude and phase response of a network from its measured transient response is described. Tables and abacs are included to facilitate numerical evaluation.

621.392.52 **1857**

Resistance and Capacitance Twin-T Filter Analysis.—L. G. Gitzendanner. (*Tele-Tech*, Feb. 1951, Vol. 10, No. 2, pp. 46–48, 71.) The twin-T circuit is redrawn in a way that makes its operation easier to understand. Formulae are developed for conditions of equal and unequal source and load impedance. Various applications in feedback circuits are discussed.

621.396.52 : 621.392.26† **1858**

U.H.F. Filters and Lenses.—Marié. (See 1842.)

621.392.54† : 621.396.662 **1859**

A High Power Attenuating Tuner for a High-Q Ten-Cm Cavity.—R. R. Perron. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, pp. 116–117.) A water-cooled attenuator constructed of polystyrene is described, which serves simultaneously to absorb about half the 900-kW peak output of a magnetron and to tune the cavity to the magnetron frequency.

621.396.611.1 **1860**

Alternative Ways in the Analysis of a Feedback Oscillator and its Application.—E. J. Post & H. F. Pit. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 169–174.) An oscillator is regarded as a closed loop comprising (*a*) an amplifier and (*b*) a passive frequency-determining network. Attenuation, introduced to limit oscillation amplitude, is often regarded as improving the phase discrimination of (*b*), but may equally well be regarded as improving the phase stability of (*a*) by negative feedback. Well-known oscillator circuits are examined from this viewpoint, and a 2-valve stabilized crystal oscillator incorporating a thermistor is analysed.

621.396.611.21 : 534.133 **1861**

Thickness-Shear and Flexural Vibrations of Crystal Plates.—R. D. Mindlin. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 316–323.) The theory of flexural motions of elastic plates, including the effects of rotatory inertia and shear, is extended to crystal plates. The equations are solved approximately for the case of rectangular plates excited by thickness-shear deformation parallel to one edge. Results of computations of resonance frequencies of rectangular AT-cut quartz plates are shown and compared with experimental data. Simple algebraic formulae are obtained relating frequency, dimensions, and crystal properties for resonances of special interest.

621.396.611.4 **1862**

The Variation of Resonance Wavelength and Damping of Cavity Resonators on Introduction of Dielectric Rings.—E. M. Philipp. (*Acta phys. austriaca*, Feb. 1948, Vol. 1, No. 3, pp. 246–258.) The problem considered is that of the circuit in which a cavity resonator is sealed directly to a valve and is arranged for continuous tuning. The influence of the location of the dielectric member forming the boundary of the evacuated region is examined. Variation of resonance wavelength and damping with position of the dielectric ring is shown in graphs. For minimum additional damping the ring should be as near as possible to the cavity boundary, i.e. in a region of low electrical field strength. The theory is confirmed by experiments.

621.396.611.4 **1863**

Phenomena in Electromagnetic Resonators near the Natural Frequency.—V. B. Shteinshleiger. (*C. R. Acad. Sci. U.R.S.S.*, 11th April 1949, Vol. 65, No. 5, pp. 669–672. In Russian.)

621.396.615 : 621.384.612.1† **1864**

The Design of Cyclotron Oscillators.—J. Backus. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, pp. 84–92.) The procedure used to design the oscillator for the 60-in. cyclotron at Berkeley is described.

621.396.645 + 621.396.727].012 **1865**

Comparative Graphical Study of Problems relating to the Cathode Follower and the Feedback Phase Inverter.—(*Radio tech. Dig., Édn franç.*, 1950, Vol. 4, No. 5, pp. 259–269.) French version of papers abstracted in 1386 of 1950 & 298 of February (Malatesta).

621.396.645.015.7 **1866**

The Problem of the 'Best' Pulse Receiver. Comparison of Various Pulse Amplifiers.—L. Huber & K. Rawer. (*Arch. elekt. Übertragung*, Nov. & Dec. 1950, Vol. 4, Nos. 11 & 12, pp. 475–484 & 523–526.) Three types of circuit are analysed, having response curves respectively peaked, multi-humped, and flat-topped. The optimum compromise between faithful reproduction and low time-lag on the one hand and high amplification and sensitivity on the other is considered in relation to the different purposes for which the receiver may be required. Four-stage amplifiers of the three types are compared, and their advantages and disadvantages summarized.

621.396.645.35 **1867**

The Automatic Compensation of Zero-Drift Errors in Direct-Coupled Feedback Systems.—F. A. Summerlin. (*Proc. Instn elect. Engrs*, Part II, Feb. 1951, Vol. 98, No. 61, pp. 59–66.) Degenerative feedback cannot correct zero-drift errors of direct-coupled amplifiers; a method is described whereby an auxiliary amplifier is used for this purpose. With suitable switching, a single auxiliary amplifier can be used to correct continuously the balance of a number of main feedback systems. The effects of the balancing circuits on stability are discussed with particular reference to a position-control servo system.

621.396.645.35 **1868**

Direct-Coupled Amplifier Starvation Circuits.—W. K. Volkers. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 126–129.) Stage gains of up to 2 500 have been obtained by using pentodes with very low screen-grid voltages and greatly increased anode loads. A 'starved' direct-coupled two-stage amplifier requires fewer components than the normal RC-coupled amplifier. The frequency response is severely limited, but can be much improved by use of negative or regenerative feedback.

621.396.645.37 : 621.396.662 **1869**

Tuning Systems employing Feedback Amplifiers.—P. G. Sulzer. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 252–258.) The effective value of an inductance or capacitance can be altered by means of a feedback amplifier of high input and low output impedance. A cathode-follower circuit can be tuned over a wide frequency range by potentiometer control. Wide-deviation frequency modulation can thus be obtained at radio frequencies; two-stage circuits have been used for resistance-tuned audio oscillators.

621.396.662.22 : 538.533 : 621.398 **1870**

The Magnetic Variometer.—A. Weis. (*Funk u. Ton*, Oct. & Nov. 1950, Vol. 4, Nos. 10 & 11, pp. 508–518 & 559–568.) From a detailed theoretical treatment of the principle of varying the inductance of a h.f. coil by magnetic means, practical design requirements for a magnetic variometer are derived, particularly in respect of the magnetic characteristics of coil and core. Ferrite cores are specially suitable. For remote tuning of receivers the dust core of the h.f. coil may be arranged in the air gap of an electromagnet. This application and the use of the device for remote control of coupling are described.

621.397.645.37 **1871**

Linearization of the Frequency Response of Wide-Band Amplifiers by Negative Feedback.—W. Dillenburger. (*Frequenz*, Jan. 1951, Vol. 5, No. 1, pp. 1–5.) The basic design of a video-frequency amplifier stage with (*a*) compensated anode circuit, (*b*) selective feedback, is considered. The latter has better response at low frequencies and low modulation depths but experiment shows that serious distortion of pulses with short rise time may occur.

GENERAL PHYSICS

534.1 + 538.56] : 621.319.55 **1872**

A New Method of Studying Relaxation Processes.—N. A. Tolstoi & P. P. Feofilov. (*Zh. eksp. teor. Fiz.*, May 1949, Vol. 19, No. 5, pp. 421–430. In Russian.)

534.232 : 538.652 **1873**

Magnetostrictive Vibration of Prolate Spheroids. Preliminary Measurements.—J. S. Kouvelites & L. W. McKeehan. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, pp. 108–111.)

535.42 : 538.56 **1874**

On Bethe's Theory of Diffraction by Small Holes.—C. J. Bouwkamp. (*Philips Res. Rep.*, Oct. 1950, Vol. 5, No. 5, pp. 321–332.) The problem is considered of the diffraction of a plane polarized e.m. wave incident normally on a small circular hole in an infinite plane conducting screen. Bethe's theory of this problem [706 of 1945; see also 1378 of 1945 (Pekeris : Bethe)] is examined and his expression for the fictitious magnetic current density in the aperture is found to be incorrect. This results in Bethe's formulae for the field near the hole being seriously in error, although his results for the distant field are correct.

537.226 **1875**

Processes in Dielectrics containing Free Charges.—B. Breyer & F. Gutmann. (*J. roy. Soc. N.S.W.*, 1949, Vol. 83, Part 1, pp. 66–74.) The free electrostatic energy, capable of doing reversible work, stored in a space-charge-filled capacitor is calculated, yielding an expression for the effective dielectric constant of a medium containing free charges. Analogies with thermal quantities are established and electrical entropy is defined. It is shown that a space-charge-filled capacitor may be loss-free if the sum of the free charges is constant or proportional to the potential difference across the plates.

537.311.33 **1876**

The Fermi Limiting Energy of Electronic Semiconductors.—H. Müser. (*Z. Naturf.*, Jan. 1950, Vol. 5a, No. 1, pp. 18–25.) The concepts of the Wilson semiconductor model are examined quantitatively, and the dependence of the Fermi limiting energy on temperature and on the nature and concentration of the impurity centres is calculated. The limiting energy is shown to be the determining factor for the electrical behaviour of the semiconductor.

537.581 **1877**

Convenient Methods for Thermionic Emission Calculations.—H. F. Ivey & C. L. Shackelford. (*Tele-Tech*, Feb. 1951, Vol. 10, No. 2, pp. 42–43, 68.) See 1902 of 1950.

538.56 : 535.312 **1878**

A Note on Reflection and Transmission.—B. A. Lengyel. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 263–264.) Propagation through the interface of two media is dependent on their reflection coefficients and transmission coefficients. Relations between these four quantities are applied to the calculation of reflection from dielectric sheets and from metal-plate structures.

538.56 : 535.312 : 621.396.677 **1879**

Reflection and Transmission at the Surface of Metal-Plate Media.—B. A. Lengyel. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 265–276.) The theory of Carlson & Heins (2756 and 3504 of 1947) has been extended; an expression is found for the reflection coefficient applicable in the presence of a diffracted beam. Tables and graphs are included for the coefficients associated with electromagnetic phenomena at the surface of metal-plate media. The magnitude of the reflection coefficient was measured at normal incidence and the phase of the transmission coefficient for angles of incidence from 0 to 25°. A satisfactory agreement with theory is obtained.

538.56 : 535.312 : 621.396.677 **1880**

Experimental Determination of the Reflection Coefficient of Metal-Plate Media.—J. Ruze & M. Young. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 277–278.) A method of measurement at 3-cm wavelength is described. Back reflections are eliminated by means of an absorbing wedge. The results obtained for structures of very thin plates are in good agreement with the theory of Carlson & Heins (2756 and 3504 of 1947). For plates of moderate thickness the measured reflection near normal incidence is higher than that predicted for an infinitely thin set of plates with identical index of refraction.

538.566 **1881**

Wave Packets, the Poynting Vector, and Energy Flow: Part 1—Non-Dissipative (Anisotropic) Homogeneous Media.—C. O. Hines. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 63–72.) The object of the present analysis is to correlate the results obtained from the two commonly used methods for finding the velocity of propagation of energy in e.m. waves, the one method involving wave packets and the other the Poynting vector. For nondissipative media the two methods are found to give the same direction of propagation of energy.

538.566.2 **1882**

The Propagation of Electromagnetic Waves in Slightly Heterogeneous Layers.—G. Eckart. (*C. R. Acad. Sci., Paris*, 28th March 1951, Vol. 232, No. 13, pp. 1294–1296.) A method developed by Schelkunoff is used to derive formulae representing the propagation of a plane wave in a system consisting of a layer whose dielectric constant varies linearly with one co-ordinate, and which is bounded along that co-ordinate by two layers with uniform dielectric constant. The formulae indicate that small departures from uniformity of the dielectric constant cannot be determined from the reflections and deformations of pulses, though large departures can be so determined.

537.122/.123 **1883**

Negative Ions. [Book Review]—H. S. W. Massey. Publishers: Cambridge University Press, New York, 2nd edn 1950, 133 pp., $2.50. (*J. Franklin Inst.*, Feb. 1951, Vol. 251, No. 2, pp. 301–302.) "A unique and satisfying combination of theory and experimental data and description is contained in this book on negative ions in the gas phase at pressures lower than a few mm of Hg."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.4 : 621.396.11 **1884**

Shortwave Radio Propagation Correlation with Planetary Positions.—Nelson. (See 1997.)

523.5 : 621.396.9 **1885**

The Interpretation of Radar Echoes from Meteor Trails.—J. Feinstein. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 37–51.) Maxwell's equations are applied to investigate the scattering of e.m. waves by a column of electrons, for various cylindrically symmetrical distributions of electron density within the column, in particular for a linear radial variation and for a uniform density distribution. Results are presented graphically in terms of the average effective dielectric constant, the ratio of cylinder circumference to wavelength, and the polarization of the incident wave. A new picture of the trail-decay process is developed from a comparison of these results with published observational data.

523.746 **1886**

Mean Areas and Heliographic Latitudes of Sunspots in the Year 1945.—(*Mon. Not. R. astr. Soc.*, 1950, Vol. 110, No. 5, pp. 501–504.)

523.746 "1950.10/.12" **1887**

Provisional Sunspot-Numbers for October to December, 1950.—M. Waldmeier. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, p. 130.)

523.75 : 550.385 **1888**

On the Expulsion of Corpuscular Streams by Solar Flares.—F. D. Kahn. (*Mon. Not. R. astr. Soc.*, 1950, Vol. 110, No. 5, pp. 477–482.) The corpuscular stream, which is the probable cause of terrestrial magnetic disturbances associated with solar flares, is thought to arise from the action of the flares themselves. The minimum momentum density of the stream required to produce a magnetic storm is calculated. Analysis of the effects of atomic resonance absorption and ionization produced by radiation pressure shows that this pressure alone is insufficient to expel such a stream from the sun.

523.75 : 550.385 **1889**

An Investigation into the Possibility of Observing Streams of Corpuscles Emitted by Solar Flares.—F. D. Kahn. (*Mon. Not. R. astr. Soc.*, 1950, Vol. 110, No. 5, pp. 483–490.)

523.752 : 537.591 **1890**

An Increase of the Primary Cosmic-Ray Intensity following a Solar Flare.—M. A. Pomerantz. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, pp. 731–733.) An increase of about 15% in the cosmic-ray intensity at altitudes of about 100 000 ft occurred about 19 hours after the commencement of the solar flare of 10th May 1949.

523.8 : 538.12 **1891**

The Strength of Interstellar Magnetic Fields.—L. Davis, Jr. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, pp. 890–891.) Discussion of the possibility of the existence of magnetic fields as strong as 10^{-4} gauss in interstellar space.

538.12 : 523.7 **1892**

Magneto-hydrostatic Fields.—S. Lundquist. (*Ark. Fys.*, 28th Dec. 1950, Vol. 2, Part 4, pp. 361–365.) Theoretical investigation of the properties of magnetic fields existing in electrically conducting fluids at rest. Such fields may have an important bearing on solar and terrestrial magnetism and may be responsible for trapping cosmic rays.

538.712 **1893**

The 1950 World Isogonic Chart.—A. M. Weber & E. B. Roberts. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 81–84.) An account of the process used by the U.S. Coast and Geodetic Survey for compiling the new edition. Punched-card machines were used. The results have been published by the U.S. Navy Hydrographic Office in one mercator and two polar charts.

550.38 "1950.07/.09" **1894**

International Data on Magnetic Disturbances, Third Quarter, 1950.—J. Bartels & J. Veldkamp. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 127–129.)

550.38 "1950.10/.12" **1895**

Cheltenham [Maryland] **Three-Hour-Range Indices *K* for October to December, 1950.**—R. R. Bodle. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, p. 130.)

550.385 "1950.07/.12" **1896**

Principal Magnetic Storms [July–Dec. 1950].—(*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 131–133.)

551.5 : 621.396.11 **1897**

Meteorological Aspects of Very-Short-Wave Propagation in the Atmosphere.—Hauer. (See 1998.)

551.510.535 **1898**

Fine Structure of the Lower Ionosphere.—R. A. Helliwell, A. J. Mallinckrodt & F. W. Kruse, Jr. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 53–62.) Night-time vertical-incidence *h't* records taken in California, at frequencies of 100 and 325 kc/s, show a series of reflections at heights between 90 and 130 km. The characteristics and polarization of the echoes are described, and several tentative explanations discussed. It is concluded that the region is divided into a series of partially reflecting strata, by a mechanism yet to be explained.

551.510.535 **1899**

Characteristics of the E_s Region at Brisbane.—R. W. E. McNicol & G. de V. Gipps. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 17–31.) The region was studied over the period 1943–1949, and the seasonal and diurnal changes are described and discussed. Two types of E_s region are distinguished, the one being formed at greater heights and descending to its final position (E_{ss}) while the other is formed in situ (E_{sc}). E_{ss} is predominant in summer, blankets strongly, and probably has a uniform ionization density; E_{sc} is a winter type, blankets little, and probably has lateral irregularities. No correlation is found between either type and sunspot number, ionospheric storms, or meteoric activity.

551.510.535 **1900**

An Investigation of Certain Properties of the Ionosphere by means of a Rapid-Frequency-Change Experiment.—B. H. Briggs. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 255–274.) Apparatus is described in which the receiver is kept in tune with the transmitter by a.f.c. circuits operated by the direct pulse from the transmitter. The tuning is corrected by each pulse and remains fixed during the intervals between pulses. A frequency range of 1 Mc/s, selected anywhere in the range 2–20 Mc/s, can be covered in 1 sec. Applications of the equipment are described and typical records are reproduced.

"The irregularities of ionization present in the normal regions are investigated by observing the irregular variations of the amplitude of the reflected wave which are produced when the frequency is altered. Irregularities of ionization are sometimes found to be localized in height. Observations of the reflection coefficient of the abnormal region E suggest that there are two distinct types of region, one an irregular region consisting of scattering clouds, the other a coherent layer with a thickness of the order of 5 km. The apparatus is also used to study the behaviour of the subsidiary critical frequencies often present below the main critical frequency of region E.

It is found that these are always decreasing whenever they are observed. Records of amplitude and group path near the critical frequency of a region can be used to determine the collision frequencies of electrons. The treatment is mainly descriptive, and a full discussion of those records which require a detailed quantitative analysis is reserved for a later paper."

551.510.535 **1901**

Vertical-Incidence Ionosphere Absorption at 150 kc/s.—A. H. Benner. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 186–190.) Recordings over a period of one year, using 150-μs pulses with a folded-dipole aerial 3 000 ft long and 96 ft high, have shown that (*a*) vertical-incidence absorption at night is of the order of 1 neper, (*b*) maximum absorption in the daytime varies from 2 nepers in midwinter to about 7 nepers (too high to measure) in midsummer, (*c*) the exponent of cos χ (χ = sun's zenith angle) that relates it to the total absorption is about 0·7 for the morning and 0·6 for the afternoon, (*d*) the exponent necessary to explain the seasonal change is considerably larger than that given by the diurnal curves, and (*e*) absorption lags behind the value of χ.

551.510.535 : 534.21 **1902**

On the Propagation of Sound over Great Distances.—Veldkamp. (See 1542 of July.)

551.510.535 : 621.396.11 **1903**

Ionosphere Review: 1950.—T. W. Bennington. (*Wireless World*, March 1951, Vol. 57, No. 3, pp. 109–111.) The variations of the monthly mean values of the F_2-layer critical frequency for Slough at noon and midnight are considered and related to corresponding variations for earlier years. The variation in relative sunspot number is similarly considered. A general indication is given of the probable trend of ionospheric conditions during 1951. The incidence during 1950 of recurrent ionospheric storms, as observed at the B.B.C. receiving station at Tatsfield, is discussed with reference to the forecasting of ionospheric conditions.

551.510.535 : 621.396.11 **1904**

The Work of the [Australian] **Radio Research Board.**—Munro. (See 1999.)

551.510.535 : 621.396.11 **1905**

The Gyro-Frequency in the Arctic E-Layer.—Scott. (See 2001.)

551.594.11 **1906**

The Semidiurnal Fluctuations of Atmospheric Potential Gradient.—H. Israel. (*Arch. Met. Geophys. Bioklimatol. A*, 4th Dec. 1948, Vol. 1, No. 2, pp. 247–251.) The double-hump curve representing the variation of atmospheric potential gradient with time of day is interpreted as consisting of two superposed curves, (*a*) a simple fluctuation with a minimum in the early morning and a maximum in the afternoon, and (*b*) a depression during the daytime, caused by turbulent exchange. The second of these components depends on the season and on the location of the observation station in respect of distance from large towns where the content of nuclei and large ions in the atmosphere is high.

551.594.221 **1907**

The Approximate Mean Height of the Thundercloud Charges taking part in a Flash to Ground.—V. Barnard. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 33–35.) Measurements are described. The results obtained from 10 flashes give an overall mean of 5·2 km, in agreement with those obtained previously by another method.

551.594.5 **1908**

Southern Extent of Aurora Borealis in North America.—C. W. Gartlein & R. K. Moore. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 85–96.) "Results are presented for the first 11 years of a study of the frequency of overhead auroras in North America as a function of latitude in a region south of the auroral zone. The data have been averaged in various ways so that monthly and annual variations are demonstrated. It appears that there is a relatively constant level of auroral activity throughout the year in the region 58° to 60° geomagnetic latitude, while auroras appearing overhead south of these latitudes are more frequent during equinoctial periods. Auroras have been seen as far south as 52° during this period every month of the year. Correlation of auroral frequency with sunspot number is not high on a month-by-month or three-month running-mean basis."

551.594.5 : 621.396.11 **1909**

A V.H.F. Propagation Phenomenon Associated with Aurora.—Moore. (See 2002.)

551.594.6 **1910**

On the Measurement and Nature of Atmospherics Produced by Electric Discharges in Snow Squalls and from Other Sources.—H. Norinder. (*Tellus*, May 1949, Vol. 1, No. 2, pp. 1–13.) The differences between atmospherics originating within snow squalls and those produced by lightning discharges are pointed out. Methods of measuring atmospherics using an open-wire aerial and c.r.o. indicator are described. Examples are given of typical meteorological situations associated with the occurrence of these atmospherics, and of typical discharge waveforms observed.

LOCATION AND AIDS TO NAVIGATION

621.396.9 **1911**

The Design and Application of a Marine Radar System.—F. W. Garrett. (*Marconi Rev.*, 1st Quarter 1951, Vol. 14, No. 100, pp. 23–38.) The trend of present practice as applicable to merchant-shipping requirements is reviewed and a detailed description is given of prototype 'Radiolocator' equipment, including particulars of the transmitter, the display console, the aerial scanning unit, motor generator, and control apparatus. Typical installations in ships are illustrated; performance is well within the Ministry of Transport specification.

621.396.9 **1912**

Harbour-Control Radar.—(*Wireless World*, March 1951, Vol. 57, No. 3, pp. 105–107.) A brief description of Decca radar Type 1. A simple corner-fed 'half-cheese' aerial is used having an aperture of 9 ft; this type of aerial shows lower side-lobe amplitudes than the conventional cheese aerial. The pulse duration used in the equipment is 0·06 μs. Examples of the high range discrimination of the system are quoted. Arrangements are incorporated for using three display consoles up to 1 000 yd from the scanner. The possibility is mentioned of using centimetre-wave links to relay information to a central office in circumstances where it is necessary to use a number of scanners to cover a harbour adequately.

621.396.9 : 523.5 **1913**

The Interpretation of Radar Echoes from Meteor Trails.—Feinstein. (See 1885.)

621.396.9 : 621.396.677 **1914**

Microwave Radar Antenna.—Chait. (See 1850.)

621.396.9.082 : 523.5 **1915**

A Circuit for Simultaneously Recording the Range, Amplitude, and Duration of Radar-Type Reflections.—

V. C. Pineo & R. C. Peck. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, p. 112.) The circuit briefly described was developed at the National Bureau of Standards for observing meteor-trail reflections of durations ranging from less than one up to several hundred seconds. A specially adapted c.r.o. display was used in conjunction with an automatic camera recorder.

621.396.932 **1916**

New Crystal-Controlled Radar Beacon.—J. W. Busby. (*Tele-Tech*, Jan. 1951, Vol. 10, No. 1, pp. 24–26, 70.) The model TB-140 'ramark' beacon, used in conjunction with marine radar receivers having a bandwidth of about 6 Mc/s, radiates on a frequency of 9·31 kMc/s, with square-wave modulation at 100, 200 or 300 kc/s. The radio frequency is derived by multiplication from a crystal-controlled master oscillator. Final multiplication is performed by a newly developed klystron valve, the SMX-32, which acts as a frequency doubler and power amplifier with an output of 2·5 W. The modulator is crystal controlled and is applied to one of the intermediate multiplying stages. Three of these units have been installed on the Ambrose Lightship outside New York harbour, the Detour Reef Lighthouse on Lake Huron, and Lightship 116 in Chesapeake Bay.

621.396.933 : 621.396.11.029.51 **1917**

Low-Frequency Radio-wave Propagation by the Ionosphere, with particular reference to Long-Distance Navigation.—Williams. (See 1761 of July.)

621.396.933.2 **1918**

Distance-Measuring Equipment for Civil Aircraft: Part 2—D.M.E. Responder Beacons.—J. P. Blom & J. D. Gilchrist. (*Proc. Instn Radio Engrs, Aust.*, Jan. 1951, Vol. 12, No. 1, pp. 9–20.) Part 1: 1392 of June (Lindsay, Blom & Gilchrist). See also 1659 of July.

621.396.933.4 : 621.396.5 **1919**

Multi-Carrier Air Communications.—(See 2016.)

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 **1920**

The Measurement of High Vacuum by Electrical Methods.—F. Wade. (*Electronic Engng*, Jan. & Feb. 1951, Vol. 23, Nos. 275 & 276, pp. 30–34 & 44–48.) A review of methods depending on (*a*) the thermal conductivity of gases, (*b*) ionization phenomena, and comparison of various types of gauge, with a table showing their optimum pressure ranges.

535.376 : 549.621.22 **1921**

Cathodoluminescence of Zinc Orthosilicate with Manganese Activator.—S. Larach & R. E. Shrader. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, p. 362.) The results of measurements of the light output, at constant current, as a function of accelerating voltage are shown for three different thicknesses of screen material. In the range investigated, 10–50 kV, equations of the form $L = KV^n$ do not fit the results, except possibly for the thickest screen (9 mg/cm^2), where n is approximately 0·8.

535.376 : 621.385.832 **1922**

The Luminescence of Cathode-Ray-Tube Screens.—R. Roulaud. (*Rev. gén. Élect.*, Feb. 1951, Vol. 60, No. 2, pp. 61–70.) The theoretical bases of luminescence in gases, liquids and solids are discussed. The luminous intensity of c.r.t. screens is calculated. Factors influencing the luminous efficiency are considered, and methods of improving efficiency and life are indicated.

535.376 : 621.397.621.2 **1923**

Saturation of Fluorescence in Television Tubes.—A. Bril & F. A. Kröger. (*Philips tech. Rev.*, Oct. 1950, Vol. 12, No. 4, pp. 120–128.) Fluorescence efficiency decreases with high beam density, due to the limited number of activator centres. For continuous bombardment, saturation depends on the activator concentration and average life-time. For discontinuous irradiation, as in television, there is no dependence on the life-time if this lies between the time for two successive passages and the time required to scan a certain spot. Sulphide phosphors are preferable for the smaller current densities, while silicates are better for conditions of heavy loading. The matching of screen-material components to avoid colour change in white-fluorescing mixtures at high current densities is discussed. Colour change may still occur with defocused beams, even for mixtures with uniform characteristics.

537.228.1 : 546.472.21 **1924**

The Piezoelectric Constant of Zinc-Sulfide.—B. D. Saksena. (*Phys. Rev.*, 15th March 1951, Vol. 81, No. 6, pp. 1012–1015.) The piezoelectric constant is calculated by a method previously used for α-quartz (2538 of 1948), for which good agreement was obtained with measurements.

537.311.32 : 539.234 : 546.23 **1925**

The Electrical Properties of Selenium Coatings.—P. H. Keck. (*J. opt. Soc. Amer.*, Jan. 1951, Vol. 41, No. 1, pp. 53–55.) Conductivity measurements were made on thin evaporated Se layers, unilluminated and illuminated. The results indicate that for coating temperatures below 50°C only vitreous Se is formed, whereas for higher coating temperatures crystalline Se is formed. From the rectifier properties of the layers it is deduced that *p*-type conduction occurs in vitreous Se and *n*-type in the crystalline material.

537.311.33 **1926**

On the Aggregation of Trapping Centers in Semiconductors or Insulators.—T. Hibi & T. Matsumura. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, pp. 884–885.)

537.311.33 **1927**

Measurement of Semiconductor Impurity Content.—C. N. Klahr & L. P. Hunter. (*Phys. Rev.*, 15th March 1951, Vol. 81, No. 6, pp. 1059–1060.) Formulae and curves are given for calculating impurity concentration from measurements of conductivity σ and Hall constant R at a low temperature (e.g., 10°–20°K for Ge) and of R at a temperature in the exhaustion range. The analysis is based on the assumption of scattering by lattice defects, by ionized impurities, and by non-ionized impurities.

537.311.33 : 546.289 **1928**

The Mobility and Life of Injected Holes and Electrons in Germanium.—J. R. Haynes & W. Shockley. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, pp. 835–843.) "The mobilities of holes injected into *n*-type germanium and of electrons injected into *p*-type germanium have been determined by measuring transit times between emitter and collector in single-crystal rods. Strong electric fields in addition to those due to injected current were employed so that spreading effects due to diffusion were reduced. The mobilities at 300°K are 1 700 cm/s per V/cm for holes and 3 600 cm/s per V/cm for electrons, with an error of probably less than 5%. The value for electrons is about 20% higher than the best estimates obtained from the conventional interpretation of the Hall effect, and the difference may be due to curved energy-band surfaces in the Brillouin zone. Studies of rates of decay indicate that recombination of holes and

electrons takes place largely on the surface of small samples with constants varying from 10^2 to $> 10^4$ cm/s for special treatments."

537.311.33 : 546.482.21 **1929**

Dependence of Electrical Properties of Cadmium Sulphide Crystals on Electrode Arrangement.—I. Broser & R. Warminsky. (*Z. Naturf.*, Jan. 1950, Vol. 5a, No. 1, pp. 62–63.) Experiments are reported which indicate the practicability of using CdS crystals for rectifiers or triodes or as photosensitive elements.

537.311.33 : 549.351.11 **1930**

On the Semiconductor Bornite.—C. W. Horton. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, p. 364.) The d.c./voltage characteristics published by El Sherbini & Yousef (2959 of 1939 and 1647 of 1941) are used to derive the energy levels for bornite between 120° and 170°C.

537.533 : 546.883 **1931**

Periodic Deviations in the Schottky Effect for Polished Tantalum.—G. B. Finn, W. B. LaBerge & E. A. Coomes. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, p. 889.) Schottky data for polished Ta wire show no significant difference from data for unpolished wire, for fields below 5×10^4 V/cm.

538.221 **1932**

Measurement of the Gyromagnetic Constant of the Ferromagnetic Elements and of some Isoelectronic Alloys of the Iron Group.—A. J. P. Meyer. (*Ann. Phys., Paris*, Jan./Feb. 1951, Vol. 6, pp. 171–210.)

538.221 **1933**

The Use of Eddy Currents for Investigating Ferromagnetic Substances.—I. Épelboin. (*Rev. gén. Élect.*, Feb. 1951, Vol. 60, No. 2, pp. 74–84.) The technique of electrolytic polishing, coupled with the study of eddy currents, has materially assisted investigations into the behaviour of inhomogeneous ferromagnetic substances subjected to weak alternating magnetic fields. The anomalous behaviour of high-permeability specimens in these fields is due partly to a nonuniform distribution of permeability with depth—the macroscopic magnetic structure of the material—and partly to a magnetic-viscosity effect associated with the diffusion of impurities. The application of successive electrolytic polishings permits the separation of these effects, but involves the destruction of the specimen. The comparative study of eddy currents provides, in a limited range of cases, a nondestructive technique for determining the structure. Technical applications are indicated and figures are quoted showing that electrolytic polishing produces a marked increase of permeability and a diminution of eddy-current losses in various standard ferromagnetic alloys. A bibliography of 41 references is appended.

538.221 **1934**

A Survey of the Possible Applications of Ferrites.—K. E. Latimer & H. B. MacDonald. (*Commun. News*, Sept. 1950, Vol. 11, No. 3, pp. 76–90.) The general mechanical and magnetic properties of ferrites, particularly those of the Mn-Zn ferrite now produced commercially as ferroxcube III, are described. Ferrites are homogeneous ceramics and may be moulded to shape and conveniently machined by grinding. The initial permeability of ferroxcube III is 700–1 500 and its resistivity 20–100 Ω.cm. Engineering applications for which ferrites have been tried, and others for which it is suggested they may be suitable, are discussed. Graphs, references and dimensioned sketches are given in illustration of the examples. The applications include filter coils of small size and high Q factor, h.f. transformers for frequencies from 10 kc/s to 50 Mc/s, power transformers for carrier frequencies, harmonic generators and delay lines.

538.652 **1935**

Magnetostriction of Fe/Pd and Ni/Pd Alloys.—Z. I. Alizade. (*C. R. Acad. Sci. U.R.S.S.*, 1st July 1950, Vol. 73, No. 1, pp. 79–81. In Russian.)

538.652 **1936**

The Magnetostriction Hysteresis of High-Coercivity Alloys.—D. I. Volkov. (*C. R. Acad. Sci. U.R.S.S.*, 1st July 1950, Vol. 73, No. 1, pp. 87–89. In Russian.)

541.183.26 : 546.78 **1937**

The Formation of Mobile and Immobile Films of Oxygen on Tungsten.—R. C. L. Bosworth. (*J. roy. Soc. N.S.W.*, 1949, Vol. 83, Part 1, pp. 31–38.) At low temperatures the condensation of O_2 on W follows Roberts' theory. At higher temperatures the process proceeds the more slowly the higher the temperature or the lower the pressure.

549.514.51 **1938**

The Use of Quartz in Radio Equipment.—C. Warenghem. (*Rev. gén. Élect.*, Feb. 1951, Vol. 60, No. 2, pp. 49–56.) An account of the properties of natural quartz and of current practice in cutting, calibrating and mounting the finished crystals. Large-scale production methods are described, and the precautions necessary to secure a high-quality product are indicated.

549.514.51 : 621.396.611.21 **1939**

Interferometric Studies on the Vibration of Piezoelectric Plates.—S. Tolansky & W. Bardsley. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 224–230.) Description of multiple-beam interference investigations. A stroboscopic method, in which the intensity of the light source is modulated at the crystal frequency, reveals the phase relations between the vibrations of different parts of the crystal surface. Typical interference patterns are illustrated.

549.514.51 : 621.396.611.21.002.2 **1940**

The Manufacture of Quartz Oscillator-Plates: Part 3—Lapping and Final Frequency Adjustment of the Blanks.—W. Parrish. (*Philips tech. Rev.*, Dec. 1950, Vol. 12, No. 6, pp. 166–177.) Description of the principles and operation of the planetary lap machine, which can deal with 30–55 plates simultaneously. The resonance frequency of the crystals can be measured during lapping by using the piezoelectric e.m.f. set up by mechanical vibration. Final frequency adjustment is made by etching in ammonium bifluoride. Part 2: 664 of March.

621.3.011.5 : 547.38 **1941**

Dielectric Properties of Ketones.—V. Daniel & K. H. Stark. (*Trans. Faraday Soc.*, Feb. 1951, Vol. 47, Part 2, pp. 149–155.) Experiments are described which prove that Müller's results published in 1937 and 1938 do not support Fröhlich's theory (3304 of 1946).

621.314.632 : 546.289 **1942**

Significance of Composition of Contact Point in Rectifying Junctions on Germanium.—W. G. Pfann. (*Phys. Rev.*, 1st March 1951, Vol. 81, No. 5, p. 882.) During forming, certain metal electrodes pass donors or acceptors to the semiconductor, modifying the properties of the junction. Donors can pass from phosphor-bronze to p-type Ge, thereby lowering the effective work function for electrons leaving the metal, and therefore improving the rectification ratio. Furthermore, the current-multiplying factor of an n-type Ge transistor increases with increasing donor (Sb) concentration in the collector electrode.

621.314.634 **1943**

A Study of Electrical Forming Phenomena at Selenium Contacts.—H. K. Henisch & J. Ewels. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 277–278.) Discussion on 926 of April.

621.315.33 : 679.57 **1944**

Wire and Cable Insulating Materials with Polyvinyl Chloride as Base.—E. A. J. Mol. (*Philips tech. Rev.*, Oct. 1950, Vol. 12, No. 4, pp. 97–110.) A comprehensive review of the mechanical, electrical and chemical properties of polyvinyl chloride insulation, with descriptions of the extrusion method of coating wire and of the principal types of 'Podur' wire.

621.315.612.4 : 546.431.82 **1945**

The Dielectric Properties of $BaTiO_3$ at Low Temperatures.—W. J. Merz. (*Phys. Rev.*, 15th March 1951, Vol. 81, No. 6, pp. 1064–1065.) Measurements on single-domain crystals from room temperature to 4·2°K are described. Dielectric loss and spontaneous polarization remain constant while permittivity decreases and the coercive field strength increases with decreasing temperature.

621.315.613.1 **1946**

Progress in Synthetic Mica.—R. D. Jackel. (*Elect. Mfg, N.Y.*, March 1950, Vol. 45, No. 3, pp. 99–103..192.) A report on investigations being made by several U.S.A. government agencies into the problem of growing, on a commercial scale, large crystals having properties comparable with those of natural mica.

621.318.3/.4].013.5 **1947**

Magnetic Field Leakage at the Edges of Airgaps.—F. Bergtold. (*Elektrotech. Z.*, 15th Feb. 1951, Vol. 72, No. 4, pp. 111–113.) The leakage field is investigated and shown graphically for a number of common cases, and an empirical formula is given for determining the extent of the associated increase of effective cross-section of the field.

621.318.323.2.042.15 **1948**

Some Properties and Tests of Magnetic Powders and Powder Cores.—C. E. Richards, P. R. Bardell, S. E. Buckley & A. C. Lynch. (*Proc. Instn elect. Engrs*, Part II, April 1950, Vol. 97, No. 56, pp. 236–245; *Elect. Commun.*, March 1951, Vol. 28, No. 1, pp. 55–69.) Conversion factors are given relating various recognized systems of expressing the electrical losses in powder cores. Test apparatus and methods used in the laboratory and in production control are described. Relevant properties of many materials are tabulated.

669.198.865 : 621.385.032.2 **1949**

Aluminum-Clad Iron for Electron Tubes.—W. Espe & E. B. Steinberg. (*Tele-Tech*, Feb. 1951, Vol. 10, No. 2, pp. 28–30, 72.) Materials developed in Germany and in the U.S.A. as substitutes for nickel are described; properties and manufacturing procedures are tabulated.

669.75 : 621.357.7 **1950**

Antimony Plate.—A. Bregman. (*Metal Progress*, Feb. 1951, Vol. 59, No. 2, pp. 245–247.) Plating techniques are described which render Sb suitable as a substitute for Ni and Cr in a variety of applications including radio and radar equipment requiring protection against corrosion.

MATHEMATICS

517.54 **1951**

An Epitomization of the Basic Theory of the Generalized Schwarz-Christoffel Transformations as used in Applied Physics.—T. J. Higgins. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 365–366.) Basic theory is outlined and 26 references are given to books and papers on these transformations, which are being increasingly used in h.f. circuit theory.

681.142 **1952**

Universal High-Speed Digital Computers: a Small-Scale Experimental Machine.—F. C. Williams, T. Kilburn & G. C. Tootill. (*Proc. Instn elect. Engrs*, Part II, Feb. 1951, Vol. 98, No. 61, pp. 13–28.) The electronic machine described uses the serial binary-digital system of number representation. Its principle of operation is explained in detail. Storage is by c.r. tube. When extended, as allowed for in the design, the machine will be intrinsically capable of performing any computation automatically.

681.142 **1953**

The General Purpose Analog Computer.—A. A. Currie. (*Bell Lab. Rec.*, March 1951, Vol. 29, No. 3, pp. 101–108.) Description of computer based on a three-stage negative-feedback amplifier, for performing addition, subtraction, multiplication, division, integration and differentiation.

681.142 **1954**

Problem Solving with the Analog Computer.—E. Lakatos. (*Bell Lab. Rec.*, March 1951, Vol. 29, No. 3, pp. 109–114.)

512.831 **1955**

Matrizen. [Book Review]—R. Zurmühl. Publishers: Springer-Verlag, Berlin, 1950, 427 pp., DM25.20. (*Frequenz*, Nov. 1950, Vol. 4, No. 11, pp. 301–302.) An introduction to matrix methods in analysis, starting from fundamental principles and including many examples of their application. ". . . can be recommended to anyone concerned with matrix calculus."

517.93 **1956**

Ordinary Nonlinear Differential Equations in Engineering and Physical Sciences. [Book Review]—N. W. McLachlan. Publishers: Clarendon Press, Oxford, 1950, 201 pp., $4.25. (*J. Franklin Inst.*, Feb. 1951, Vol. 251, No. 2, p. 303.) "The book is *not* an analytical treatise with technical applications. It aims to show how certain types of nonlinear problems may be solved, and how experimental results may be interpreted by aid of nonlinear analysis. The contents of the book include: General Introduction, Equations Readily Integrable, Equations Integrable by Elliptic Integrals and Functions, Equations having Periodic Solutions, Method of Slowly Varying Amplitude and Phase, The Equivalent Linear Equation, Equations having Periodic Coefficients, and Graphical and Numerical Solutions."

518.12 + 517.392 **1957**

Numerical Mathematical Analysis. [Book Review]—J. B. Scarborough. Publishers: Johns Hopkins Press, Baltimore, 2nd edn 1950, 511 pp., $6.00. (*J. Franklin Inst.*, Feb. 1951, Vol. 251, No. 2, pp. 303–304.) "Chapters have been included on the numerical solution of partial differential equations and on integral equations. The chapters on the numerical solution of ordinary differential equations and on numerical integration have been thoroughly revised and enlarged."

MEASUREMENTS AND TEST GEAR

621.317.029.6 : 061.3 **1958**

Conference on High-Frequency Measurements, Washington, D.C., 10th–12th January 1951.—(*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 208–211.) Summaries are given of the following technical papers presented:

1. Program for Atomic Frequency and Time Standards —A Survey.—H. Lyons.
2. Improved N.B.S. Ammonia Clock.—B. F. Husten.
3. The Stabilization of a Microwave Oscillator with an Ammonia Absorption Line Reference.—E. W. Fletcher & S. P. Cooke.

4. Performance of Oscillators Frequency-Controlled by Gas Absorption Lines.—L. E. Norton.
5. Millimeter-Wave Measurements.—W. Gordy.
6. Quartz-Crystal Frequency Standards.—W. D. George.
7. High-Frequency Crystal Units for Primary Frequency Standards.—A. W. Warner.
8. Inflection-Point Method of Measuring Q at Very-High Frequencies.—N. E. Beverly.
9. A Precise Sweep-Frequency Method of Vector Impedance Measurement.—D. A. Alsberg.
10. Precision Coaxial Resonance Line for Impedance Measurements.—H. E. Sorrows, R. E. Hamilton, W. E. Ryan & Ming S. Wong.
11. A 2600- to 4000-Mc/s VSWR-Measuring Set.—S. F. Kaisel & J. W. Kearney.
12. Measurement of Waveguide and Coaxial-Line Impedances with a Circular Waveguide.—A. E. Laemmel.
13. Survey of Microwave Dielectric Techniques for Small Liquid and Solid Samples.—G. Birnbaum.
14. Microwave Spectroscopy with Application to Chemistry, Nuclear Physics, and Frequency Standards.—L. J. Rueger, R. G. Nuckolls & H. Lyons.
15. Recording Atmospheric Index of Refraction at Microwaves.—G. Birnbaum, S. J. Kryder & R. R. Larson.
16. Measurement of Microwave Field Patterns using Photographic Techniques.—W. E. Kock.
17. Absolute Microwave Power Measurements.—A. C. Macpherson & D. M. Kerns.
18. Broad-Band Bolometer Development.—W. E. Waller.
19. Calibrating Ammeters above 100 Mc/s.—H. R. Meahl & C. C. Allen.
20. A Microwave Oscillograph.—W. B. Sell & J. V. Lebacqz.
21. Precision Millidecibel Waveguide-Attenuation Measurements.—J. H. Vogelman.
22. Dissipative and Piston Attenuator Corrections.—C. M. Allred.
23. A Field-Strength Meter for 600 Mc/s.—J. A. Saxton.
24. Measuring Techniques for Broad-Band Long-Distance Radio Relays.—W. J. Albersheim.
25. Wide-Band Swept-Frequency Measurements applicable to Traveling-Wave Tubes.—F. E. Radcliffe.
26. Microwave Techniques in the 28000- to 300000-Mc/s Region.—L. Swern.
27. Measurement of Characteristics of Crystal Units.—L. F. Koerner.
28. Reflecting Surface to Simulate an Infinite Conducting Plane at Microwave Frequencies.—S. J. Raff.

621.317.3.001.4 : 621.396.615.142.2 **1959**

3-cm Low-Voltage Reflex Klystron, Type RHK 6332.—Chantereau, Musson-Genon & Métivier. (See 2069.)

621.317.335 **1960**

The Double-Superheterodyne Principle applied to Dielectric Measurements.—E. B. Baker. (*Rev. sci. Instrum.*, Jan. 1951, Vol. 22, No. 1, pp. 34–36.) Description, with circuit diagrams, of generator-detector equipment for the range 0–16 kc/s, which includes two mixers and a common variable-frequency local oscillator. Advantages are (*a*) single-dial tuning, (*b*) freedom from relative frequency drift and (*c*) improved signal/noise ratio. The equipment is particularly suitable for use with automatic-balancing bridges. Extension of the range to 600 kc/s is practicable.

621.317.335.2.088 **1961**

The Influence of Mutual and Self Inductances on the Accurate Determination of Capacities.—A. Michels & C. A. ten Seldam. (*Appl. sci. Res.*, 1951, Vol. B2, No. 2, pp. 73–76.) Formulae indicating the effect of lead inductances are presented. To avoid measurement errors, common leads should be avoided.

621.317.336/.34.029.6 **1962**

Application of Directional Couplers to Amplitude-Ratio and Phase Measurement at Very High Frequencies.—B. M. Sosin. (*Marconi Rev.*, 1st Quarter 1951, Vol. 14, No. 100, pp. 39–57.) A critical survey is made of methods of accurate measurement of impedance or admittance at v.h.f., and a new technique is described in which two waves are sampled independently by directional couplers and compared in amplitude and phase. Essential parts of the apparatus are discussed and procedures are outlined for input impedance measurement, matching of input impedance, matching of discontinuities, and for measurements of insertion-loss, crosstalk, power division at a junction, and attenuation constant of a transmission line. The method of setting up the equipment is described and an estimate made of the probable errors.

621.317.336.088 **1963**

Evaluation of Coaxial-Slotted-Line Impedance Measurements.—H. E. Sorrows, W. E. Ryan & R. C. Ellenwood. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 162–168.) The errors introduced by probe coupling, attenuation and by the slot can be estimated and those due to structural defects can be measured. The errors in measuring the voltage s.w.r. and the position of the nodes are analysed, and the accuracy of the resultant impedance calculation is determined.

621.317.351 **1964**

The Radiospectroscope SFM 101.—W. C. van B. Jutting. (*Commun. News*, Dec. 1950, Vol. 11, No. 4, pp. 120–125.) The apparatus comprises f.m. oscillator, sawtooth oscillator, mixer, amplifiers, detector and c.r.o., and is adapted to work in conjunction with any receiver having an i.f. between 400 and 500 kc/s. The observation frequency band is $\pm$ 50 kc/s with respect to a centre frequency of 140 kc/s + receiver i.f. Application is particularly for measurement of frequency deviation, carrier shift and modulation distortion in f.m. transmitters, and of modulation depth and parasitic f.m. and asymmetrical modulation in a.m. transmitters.

621.317.361 **1965**

A Modern Frequency Measuring Installation.—H. A. Taylor & E. C. Rundquist. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 98–102.) An account of R.C.A. equipment and methods providing, through receiving stations at Riverhead, Long Island, Point Reyes, California, and Manila, a commercial service with world-wide coverage except at v.h.f. and u.h.f., where the effective range is limited to 200–500 miles, depending on propagation conditions. The primary frequency standards are 100-kc/s quartz crystal oscillators, which are frequently compared with observatory time signals and with standard-frequency radio transmissions. The main equipment measures frequencies over the range from 100 kc/s to 26 Mc/s, with extensions to provide measurements from 15 kc/s to 100 kc/s and from 26 Mc/s to 500 Mc/s. The average deviation of the primary standard-frequency source is approximately 1 cycle in 5×10^8, while the possible error in the interpolation system is less than 2 c/s at any frequency from 15 kc/s to 26 Mc/s. The aerial arrays permit a wide choice of directive pattern, so that interference can be largely eliminated.

621.317.7 : 621.392.26† **1966**

Some Aspects of Waveguide Technique.—A. E. Pannenborg. (*Commun. News*, Sept. 1950, Vol. 11, No. 3, pp. 65–75.) See 946 of April.

621.317.723 **1967**

Circuits and Tubes for Ultra-Sensitive Electrometers.—F. E. O'Meara. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, pp. 106–108.)

621.317.725 : 621.396.622.7 **1968**

The Audion Valve Voltmeter.—H. Boucke. (*Frequenz*, Nov. 1950, Vol. 4, No. 11, pp. 281–289.) Discussion of the design details of the instrument, particularly component values in the detector circuit, methods of calibration and adjustment, and stabilization. The circuit comprises basically a grid-leak detector and a compensating valve in two arms of a bridge circuit. Methods are described for extending the range.

621.317.727.027.213 **1969**

High-Sensitivity Automatic Recording Potentiometer.—S. Ekelöf & B. Nilsson. (*IVA, Stockholm*, 1950, Vol. 21, No. 4, pp. 153–162.) The arrangement includes a standard potentiometer circuit and an automatically balancing feedback circuit consisting of null galvanometer, photocell and two-stage d.c. amplifier. Voltage variations $< 0{\cdot}001\%$ can easily be recorded. Accuracy is to within 5–10% of full-scale deflection. Possible applications include investigation of the stability of d.c. voltage sources such as primary batteries.

621.317.79 : 621.3.018.78† : 621.395.623.7 **1970**

An Automatic Nonlinear Distortion Analyzer [for loudspeakers].—Olson & Pennie. (See 1836.)

621.317.79 : 621.396.619 **1971**

The N.R.U. [Nederlandse Radio Unie] **Modulation Monitor Type ON301.**—R. Y. Drost. (*Tijdschr. ned. Radiogenoot.*, Jan. 1951, Vol. 16, No. 1, pp. 1–11. Discussion, p. 12.) The unit described forms part of the equipment standardized for use in stations of the Netherlands Union of Broadcasting Corporations. The scale of the instrument is logarithmic for modulation depths up to 50% and linear above that value. Pulses of amplitude within the linear scale range and of duration below a given time interval (e.g. 500 ms) are automatically lengthened to enable their value to be correctly indicated. Light-spot indicators are used.

621.396.619.13 : [621.317.761 + 621.317.7 : 621.396.619 **1972**

Communication-System Monitor.—J. E. Stiles. (*FM-TV*, Feb. 1951, Vol. 11, No. 2, pp. 16–18.) General description, with block diagram and simplified circuit diagrams of the discriminator and reference oscillator, of the Hewlett-Packard f.m. monitor, Model 337, which enables measurements to be made of both carrier frequency and peak modulation swing of transmissions in the range 30–175 Mc/s.

621.317 **1973**

Basic Electrical Measurements. [Book Review]—M. B. Stout. Publishers: Prentice-Hall, New York, 1950, 504 pp., $5.75. (*J. Franklin Inst.*, Feb. 1951, Vol. 251, No. 2, p. 304.) "The book would make a fine text for an electrical engineering course."

621.317.323 **1974**

High-Frequency Voltage Measurements. National Bureau of Standards Circular 481. [Book Review]—M. C. Selby. Publishers: U.S. Department of Commerce, 14 pp., $0.20. (*Tijdschr. ned. Radiogenoot.*, Jan. 1951, Vol. 16, No. 1, p. 54.) Principles and methods are given for both high-accuracy (to within 1%) and medium-accuracy (to within 5%) measurements. The frequency range covered is from about 10 kc/s to about 100 Mc/s. A useful list of references is included.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534 : 577 **1975**

A Bibliography on Sonic and Ultrasonic Vibration: Biological, Biochemical and Biophysical Applications.—Naimark, Klair & Mosher. (See 1813.)

534.321.9.001.8 **1976**

Latest Developments in Ultrasonics.—A. R. Laufer. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 82–86.) Discussion of the generation and use of ultrasonic energy, with particular reference to the high-power generator in use at the University of Missouri. Some methods of mounting quartz crystals in ultrasonic oscillators are illustrated. A short-range optical communication system, in which the modulator is a xylene-filled ultrasonic-diffraction cell, can be used for distances up to 5000 yd at night with an infrared light filter.

537.533 **1977**

On the Spherical Aberration of Electron Emission Systems.—O. Klemperer & Y. Klinger. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 231–233.) Both the spherical aberration and the apparent source diameter for a 'hairpin' emission system are found to be less than the corresponding values for various other emission systems.

621.317.083.4 **1978**

Radio Telemetry.—M. H. Nichols & L. L. Rauch. (*Rev. sci. Instrum.*, Jan. 1951, Vol. 22, No. 1, pp. 1–29.) An extensive theoretical survey of possible methods and discussions of characteristics. Frequency and time-division types are considered in detail and comparisons of crosstalk and fluctuation noise, complexity and reliability are made. Recording and instrumentation are considered, and crosstalk due to overload and restricted bandwidth are calculated. Information theory is applied and the efficiency of various systems estimated.

621.38.001.8 **1979**

Detecting Tramp Metal in Logs and Iron Ore.—C. W. Clapp. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 88–93.) Based on 1950 National Electronics Conference paper. Analysis of the problem of detecting a small metal sphere in an a.c. field, and a description of practical application to the detection of (*a*) bullets and shell fragments in logs floating down a flume, (*b*) broken drills in highly magnetic taconite iron ore carried on a conveyor belt.

621.38.001.8 **1980**

Microwaves offer New Control Functions.—W. C. White. (*Elect. Mfg, N.Y.*, March 1950, Vol. 45, No. 3, pp. 86–89, 170.) A general account is given of methods suitable for radiating and detecting microwaves, and the particular conditions under which microwave relays can usefully supplant photocell relays are discussed.

621.384.612.1† : 621.396.615 **1981**

The Design of Cyclotron Oscillators.—J. Backus. (*Rev. sci. Instrum.*, Feb. 1951, Vol. 22, No. 2, pp. 84–92.) The procedure used to design the oscillator for the 60-in. cyclotron at Berkeley is described.

621.384.612.2† **1982**

An Air-Cored Synchro-Cyclotron for 400-MeV Protons.—L. Riddiford. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 218–224.)

621.385.38 : 621.316.7.076.7 **1983**

Thyratron Grid-Circuit Design.—J. H. Burnett. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 106–111.) Seven basic methods for phase control of firing in thyratrons, and four primary sources of noise voltages in thyratron circuits are analysed, with circuit diagrams and waveforms. Criteria for reliable operation are based on consideration of maximum critical grid current, transients and frequency.

621.385.833 **1984**

Investigation of Electron Optical Properties of an Electrostatic Focusing System.—W. E. Spear. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 233–243.) Paraxial ray tracing together with a dynamical determination of electron trajectories in the cathode region leads to a complete analysis of electron-optical immersion systems. The method is applied to an electrode arrangement for use in a fine-focus X-ray tube [1450 of June (Ehrenberg & Spear)].

621.385.833 **1985**

Electron-Optical Systems with Helical Axis.—D. Gabor. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 244–255.) Systems with curved axes can be designed which are equivalent to centred systems in the Gaussian approximation, but have chromatic error of opposite sign and thus can be used as correcting elements. Theory is given of such systems with a helical axis, with a helical e.s. field shaped so as to eliminate first-order astigmatism. The properties of such systems, and their applications, are discussed.

621.385.833 **1986**

Electron Microscopes.—M. Locquin. (*Électronique, Paris*, Oct. 1950, No. 47, pp. 5–21, 44–49.) A comprehensive illustrated survey of models available on the world market, compiled from information supplied by the makers.

621.385.833 **1987**

International Congress on Electron Microscopy.—(*Électronique, Paris*, Oct. 1950, No. 47, pp. 22–38.) Summaries are given in French and English of papers presented at the congress held in Paris, from 14th to 22nd September 1950.

621.385.833 : 621.396.615.141.2 **1988**

Electron Optical Exploration of Space Charge in a Cut-Off Magnetron.—D. L. Reverdin. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 257–262.) Detailed account of the work noted in 2315 of 1950.

621.387.4† **1989**

The Elimination of the End Effects in Counters.—A. L. Cockroft & S. C. Curran. (*Rev. sci. Instrum.*, Jan. 1951, Vol. 22, No. 1, pp. 37–42.)

621.387.4† **1990**

A Note on the Successive Peaks of Spurious Counts in G-M Tubes.—J. E. Kupperian, Jr, P. C. Murray & H. Feeny. (*Rev. sci. Instrum.*, Jan. 1951, Vol. 22, No.1, pp. 60–61.)

621.396.662.22 : 538.533 : 621.398 **1991**

The Magnetic Variometer.—Weis. (See 1870.)

621.385.833 **1992**

The Electron Microscope, its Development, Present Performance and Future Possibilities. [Book Review]—D. Gabor. Publishers: Chemical Publishing Co., New York, 2nd edn 1948, 164 pp., $4.95. (*J. opt. Soc. Amer.*, Dec. 1950, Vol. 40, No. 12, pp. 879–880.) Written from the point of view of the physicist; mathematical details are kept to a minimum.

PROPAGATION OF WAVES

538.566 **1993**

Wave Packets, the Poynting Vector, and Energy Flow: Part 1 — Non-Dissipative (Anisotropic) Homogeneous Media.—Hines. (See 1881.)

538.566.2 **1994**

The Propagation of Electromagnetic Waves in Slightly Heterogeneous Layers.—Eckart. (See 1882.)

621.396.11 **1995**

Fluctuations of the Angle of Departure of Ionospheric Waves.—W. Budde. (*Arch. elekt. Übertragung*, Dec. 1950, Vol. 4, No. 12, pp. 509–516.) A method is described which enables the vertical and azimuthal components of the angle of departure to be determined in one step. The equipment comprises in principle two pairs of transmitting aerials whose base lines, 2λ long, are set at a given angle. The total phase difference D between the radiations from the aerials of a single pair is the sum of the difference due to the geometrical separation, δ, and the electrical phase difference, Δ. δ is kept constant while Δ is varied until the field strength recorded at the receiver is zero, i.e. $D = \pi$; δ is thus determined. The direction of departure is found from the phase differences for the two pairs of aerials. For a transmission distance of 1 500 km fluctuations of elevation angle of 5°–10° were measured. The deviations from great circle ranged from 3° to 5°. Operating frequency was 9·8 Mc/s.

621.396.11 **1996**

World Charts in Short-Wave Engineering.—T. W. Bennington. (*B.B.C. Quart.*, Spring 1951, Vol. 6, No. 1, pp. 43–54.) The importance of the ionosphere in making possible long-distance communication is discussed. Advance information on ionospheric conditions several months ahead is desirable to enable frequency allocations to be used efficiently. A qualitative description is given of the methods used in deriving the maximum usable frequency for propagation via the ionosphere. The effects of ionospheric absorption and noise at the receiver site on the lowest usable frequency are described.

621.396.11 : 523.4 **1997**

Shortwave Radio Propagation Correlation with Planetary Positions.—J. H. Nelson. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 26–34.) Consideration is given to 'planetary configurations' in which the radius vectors of two planets include an angle equal to zero or a multiple of 90°. The distribution in time of such configurations was compared with observations of disturbances in s.w. propagation conditions over the N. Atlantic; good correlation is claimed.

621.396.11 : 551.5 **1998**

Meteorological Aspects of Very-Short-Wave Propagation in the Atmosphere.—A. Hauer. (*Tijdschr. ned. Radiogenoot.*, Jan. 1951, Vol. 16, No. 1, pp. 39–52.) An outline is given of the theory of refraction and scattering of radio waves in the troposphere and of the influence of humidity and temperature gradients on propagation. The criterion for a ray to be refracted back to earth is discussed. Metre waves can be received beyond the horizon even when the vertical gradient of modified refractive index is not negative; when this gradient is negative, ducts are formed and superrefraction occurs. Measurements made over the Caribbean Sea, on a wavelength of 9 cm, are compared with calculated values of signal strength as a function of distance from the transmitter, and of the scattering due to 4-, 6- and 8-cm atmospheric 'blobs'. Agreement is good. A comparison also indicates some correlation between radiosonde measurements and the observed quality of certain 3-m communication links in Holland.

621.396.11 : 551.510.535 **1999**

The Work of the [Australian] **Radio Research Board.**—G. H. Munro. (*Proc. Instn Radio Engrs, Aust.*, Feb. 1951, Vol. 12, No. 2, pp. 41–43.) The history of the Board is reviewed and the development of the ionospheric prediction service during the war period 1939–1946 is described. Post-war investigations of movements in the ionosphere F layer are surveyed, and other investigations are discussed briefly.

621.396.11 : 551.510.535 **2000**

Ionosphere Review: 1950.—Bennington. (See 1903.)

621.396.11 : 551.510.535 **2001**

The Gyro-Frequency in the Arctic E-Layer.—J. C. W. Scott. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 1–16.) Calculation of the gyrofrequency in the E layer in Northern Canada from measurements of critical-frequency differences gives a lower value of terrestrial magnetic field than extrapolation from ground-level values. At one station a large semidiurnal variation with maxima at 0600 and 1800 L.M.T. was found. The explanation given for similar effects in the F layer (1471 of June) is not applicable. In the E layer the effects may be caused by a variable concentration of heavy ions, rising to over 4 000 times the density of free electrons.

621.396.11 : 551.594.5 **2002**

A V.H.F. Propagation Phenomenon associated with Aurora.—R. K. Moore. (*J. geophys. Res.*, March 1951, Vol. 56, No. 1, pp. 97–106.) Abnormal propagation has been observed at frequencies between 28 and 148 Mc/s during auroral activity. The effects observed include a very high rate of fading, such as to render radio-transmitted speech unintelligible, the absence of any skip effect, and the necessity for the transmitting and receiving aerials to be pointed northwards. The correlation of these effects with the southerly limit of auroral displays is examined.

RECEPTION

621.316.726 : 621.396.62 **2003**

Automatic Frequency Control for Single-Sideband Receivers.—J. L. Arends. (*Commun. News*, Dec. 1950, Vol. 11, No. 4, pp. 101–119.) A brief explanation is given of s.s.b. operation, and the elements of an a.f.c. circuit are discussed. Two forms of a.f.c. are distinguished, viz. electronic and electromechanical; a combination of the two is required for s.s.b. reception during fading. A description is given of an electromechanical system using the 'flotor'—a small variable capacitor comprising a tube of insulating material containing oil, within which a metal core is able to move freely. Special problems arising when a.f.c. is used in conjunction with frequency-shift telegraphy are mentioned.

621.396.621 **2004**

Communication Receiver Design.—D. Heightman. (*Radiotronics*, Oct. 1950, No. 145, pp. 99–105, 121.) Reprint. See 1235 of 1950.

621.396.645.015.7 **2005**

The Problem of the 'Best' Pulse Receiver. Comparison of Various Pulse Amplifiers.—Huber & Rawer. (See 1866.)

621.396/.397.82 **2006**

Open-Field Test Facilities for Measurement of Incidental Receiver Radiation.—C. G. Seright. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 45–52.) Description of a test setup for measuring interference radiation from television and f.m. receivers.

621.397.8 **2007**

Investigation of Ultra-High-Frequency Television Transmission and Reception in the Bridgeport, Connecticut, Area.—Guy. (See 2037.)

621.396.62.004.67 : 621.397.62 **2008**

Wireless Servicing Manual. [Book Review]—W. T. Cocking. Publishers: Iliffe & Sons, London, 8th edn 1950, 296 pp., 12s. 6d. (*J. Brit. Instn Radio Engrs*, Feb. 1951, Vol. 11, No. 2, p. vi.) Revised and brought up to date; the chapter on television has been completely rewritten. Problems of hum and distortion are treated very thoroughly.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 **2009**

Maxwell's Demon cannot operate: Information and Entropy: Part 1.—L. Brillouin. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 334–337.) "In an enclosure at constant temperature, the radiation is that of a 'black body', and the demon cannot see the molecules. Hence, he cannot operate the trap door and is unable to violate the second principle. If we introduce a source of light, the demon can see the molecules, but the over-all balance of entropy is positive. This leads to the consideration of a cycle

Negentropy → Information → Negentropy

for Maxwell's demon as well as for the scientist in his laboratory. Boltzmann's constant k is shown to represent the smallest possible amount of negative entropy [negentropy] required in an observation."

621.39.001.11 **2010**

Physical Entropy and Information: Part 2.—L. Brillouin. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 338–343.) "The laws of statistical thermodynamics are used for the definition of entropy, and it is shown that the definition of information can be reduced to a problem of Fermi-Dirac statistics or to a generalized Fermi statistics. With these definitions, the entropy of a certain message can be defined, and the information contained in the message can be directly connected with the decrease of entropy in the system.

This definition leads directly to the formulas proposed by C. E. Shannon [455 of February (Shannon & Weaver)] for the measure of information, and shows that Shannon's 'entropy of information' corresponds to an equal amount of negative entropy in the physical system. The physical background of the whole method is discussed and found in agreement with previous discussions."

621.39.001.11 : 061.3 **2011**

Survey of the Work of the Conference on Information Theory.—M. D. Indjoudjian. (*Ann. Télécommun.*, Jan. 1951, Vol. 6, No. 1, pp. 27–31.) An account of the proceedings at the symposium held in London in September 1950. An account in English was noted in 984 of April (Jackson).

621.39.001.11 : 621.397.5 **2012**

Television and the Transmission of Information.—Delbord. (See 2028.)

621.395.44 : 621.396.619.2 **2013**

A 48-Channel Carrier Telephone System: Part 2—Apparatus Design.—G. H. Bast, D. Goedhart & J. F. Schouten. (*Commun. News*, Sept. 1950, Vol. 11, No. 3, pp. 91–100.) Reprint. See 3251 of 1949. Reprint of part 1 noted in 739 of March.

621.396.216 : 621.396.619.24 **2014**

The Single-Sideband System of High Frequency Radio Transmission.—E. T. Wrathall & C. P. Beanland.

(*Marconi Rev.*, 1st Quarter 1951, Vol. 14, No. 100, pp. 2–22.) A simple and general treatment of the subject is given. The theoretical advantages of a s.s.b. system over d.s.b. systems, considered from the signal/noise, bandwidth and carrier-suppression viewpoints, are shown to be borne out in practice. A comparison of radiated-power requirements and d.c. anode input powers shows a marked economy in favour of the s.s.b. transmission. Five ways of using the sideband spectrum for music, speech and telegraph channels are outlined, and the power distribution per channel is considered for each case.

Diagrams of typical s.s.b. generating equipment and monitoring circuits for carrier level, distortion and interchannel crosstalk are given. Details of the power amplifier circuits of the transmitter are discussed. The process of multiple heterodyning in a typical s.s.b. receiver to produce sidebands related to a nominal carrier, and their demodulation and separation, are described. The benefits of using spaced-aerial reception, especially when receiving telegraph signals on a s.s.b. system, are explained.

621.396.5 : 621.395.632 **2015**

The Auto-Call.—R. V. Anderson. (*CQ*, Feb. 1951, Vol. 7, No. 2, pp. 33–35.) Description of a device which can be fitted to a standard receiver to enable a station not 'on the air' to be called from any mobile station. The caller whistles a sequence of pulses into his microphone. At the receiver these pulses operate an anode-circuit relay controlling a stepping switch which connects the loudspeaker to the output valve, at the same time cutting out a high-pass audio filter inserted to prevent false operation by speech modulation. Conversation can then be carried on as usual.

621.396.5 : 621.396.933.4 **2016**

Multi-Carrier Air Communications.—(*Wireless World*, March 1951, Vol. 57, No. 3, pp. 92–94.) A brief description of a v.h.f. system of radiotelephony for the control of civil aircraft, with particular reference to arrangements in the southern part of England. To ensure adequate coverage, five combined transmitting and receiving stations are dispersed over the area, all being operated from a common control point. Outline particulars of the transmitting and receiving equipment are given and the arrangements by which more than one receiver or more than one transmitter can be used with a single aerial are described. The scheme for frequency allocation is indicated. See also *Electronic Engng*, March 1951, Vol. 23, No. 277, pp. 86–91.

621.396.619 : 621.391.1 **2017**

The Assessment of Modulation Systems using the Communication-Theory Concept of Channel Capacity.—R. Piloty. (*Arch. elekt. Übertragung*, Dec. 1950, Vol. 4, No. 12, pp. 493–508.) Channel capacity is defined in terms of the highest possible rate of transmission of single-valued signals of a given duration. Formulae are developed giving the capacity for both continuous and discrete channels by a method less rigorous than that of Shannon (1361 and 1649 of 1949). The theory is used to compare the frequency-band and signal/noise-ratio requirements for various modulation systems for the same channel capacity; results are shown in graphs. Pulse code modulation is shown to be advantageous from this point of view.

621.396.65 **2018**

Emergency Wire-Line to Radio Circuits.—D. Talley. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 195–215.) Description of the construction and operation of equipment for temporary connection of a radio circuit, which is essentially a 4-wire system, to a 2-wire telephony network.

621.396.712 **2019**

B.B.C. Stations.—(*Wireless World*, March 1951, Vol. 57, No. 3, p. 104.) An up-to-date list of B.B.C. medium- and long-wave stations, stating the programme transmitted and the wavelength and power used.

SUBSIDIARY APPARATUS

621-526 **2020**

The Position Synchronization of a Rotating Drum.—F. C. Williams & J. C. West. (*Proc. Instn elect. Engrs*, Part II, Feb. 1951, Vol. 98, No. 61, pp. 29–34.) Description of a servo system, using a crystal reference oscillator, for controlling within fine limits the angular speed and angular position of a drum.

621-526 **2021**

Ascertaining the Critical Servo Gain.—A. Schlang. (*Proc. Inst. Radio Engrs*, March 1951, Vol. 39, No. 3, p. 290.) A simplified method is described for determining, from the system determinantal equation, the critical gain resulting in instability.

621-526 : 659.25.001.11 **2022**

A General Scheme relating to a Problem in Cybernetics.—S. Colombo. (*C. R. Acad. Sci., Paris*, 28th March 1951, Vol. 232, No. 13, pp. 1287–1288.) General formulae are given representing the operation of the system comprising apparatus to be regulated, comparator-discriminator, and servo-control mechanism.

621-526 : 659.25.001.11 **2023**

Reduction of a Problem in Cybernetics to a Problem of 'Pursuit' in a Hilbert Space.—R. Vallée. (*C. R. Acad. Sci., Paris*, 28th March 1951, Vol. 232, No. 13, pp. 1288–1290.) Development of a particular method of considering the problem dealt with in 2022 above.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 **2024**

Perspective Distortion in TV Pictures.—E. C. Lloyd. (*TV Engng, N.Y.*, Jan. & Feb. 1951, Vol. 2, Nos. 1 & 2, pp. 12–15 & 18–22.) 'Proper viewing distance' is defined; the perspective distortion introduced at other viewing distances is evaluated quantitatively, and methods are developed for minimizing this distortion and for assessing the adequacy of standards of picture definition.

621.397.5 **2025**

The Sharpness of the Television Picture when the Picture Element has a Bell-Shaped Illumination-Intensity-Distribution Curve.—E. Schwartz. (*Arch. elekt. Übertragung*, Dec. 1950, Vol. 4, No. 12, pp. 517–522.) Theoretical investigation of the form of the photocell-current curve obtained when a circular spot, across which the electron density decreases from centre to periphery, is scanned across a line where the picture brightness changes suddenly. A formula is developed in the form of a power series, of which the independent variable is proportional to the distance between the centre of the scanning spot and the brightness discontinuity. The resulting curve is S-shaped, and its slope on crossing the discontinuity is greater than that of the straight line corresponding to a uniform-intensity square scanning spot. Thus for the same sharpness the circular spot may be larger.

621.397.5 : 535.623/.624 **2026**

Colour Television.—M. Alixant. (*Radio tech. Dig., Édn franç.*, 1950, Vol. 4, Nos. 4 & 5, pp. 221–243 & 271–274. Bibliography, pp. 275–278.) Review with illus-

trated descriptions of the features of different transmission and reception systems, particularly American, classified according to principle of operation.

621.397.5 : 535.623 **2027**

Analysis of Synchronizing Systems for Dot-Interlaced Color Television.—T. S. George. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 124–131.) A mathematical analysis of random phase error due to noise and of static phase error caused by variations in frequency from crystal to crystal in two systems which might be used to synchronize the receiver dot or sampling frequency, viz:—(*a*) a simple high-*Q* resonant filter, and (*b*) an oscillator with a.f.c. Calculations made for i.f. carrier/noise ratios of 1, 3 and 5 (the critical range) show that for a phase-error tolerance of 10° receivers operating without manual control of the dot frequency could probably be designed for carrier/noise ratios down to about 2.

621.397.5 : 621.39.001.11 **2028**

Television and the Transmission of Information.—Y. Delbord. (*Ann. Télécommun.*, Jan. 1951, Vol. 6, No. 1, pp. 11–22.) Paper given at the symposium on Signal and Information Theory, Paris, April/May 1950. Television is compared with various other ways of transmitting information, such as telegraphy, telephony, teleprinting, etc. The comparison is relatively easy for still material, and indicates that all the systems make approximately equally good use of the frequency channel. For a moving picture the assessment is more difficult; the frequency band required in this case is 1 000–2 000 times that for transmitting a still picture. Methods both old and new for reducing this factor are considered, e.g., line interlacing, division of the vision frequency spectrum into sub-bands transmitted sequentially, use of receiver c.r. tube with long-persistence screen. Reduction of the frequency spectrum would make possible the use of lower carrier frequencies and hence a great extension of television.

621.397.5 : 778.1 **2029**

Some Factors in Pictorial Reproduction Processes with Special Reference to Television.—R. G. Hopkinson, R. B. Mackenzie & R. D. Nixon. (*Photogr. J.*, Jan./Feb. 1951, Vol. 91B, No. 1, pp. 2–10.) The factors which determine the quality of pictorial reproduction in general are discussed, and their application in television practice is indicated.

621.397.5(73) **2030**

Television Broadcasting in the United States, 1927-1950.—D. G. Fink. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 116–123.) A review of the development of transmission methods and equipment, receivers, and theatre and colour television.

621.397.6 : 621.396.67 **2031**

TV Receiving Antenna Research, Design and Production.—R. G. Peters. (*TV Engng, N.Y.*, Feb. 1951, Vol. 2, No. 2, pp. 8–10.) A report of methods used by some U.S. manufacturers.

621.397.62 **2032**

P-M [permanent-magnet] **Focus Devices for Picture Tubes.**—K. James & R. T. Capodanno. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 94–97.) A discussion of the requirements of a good permanent-magnet focusing system, and of the various imperfections often occurring in practice. The stray field must be reduced to a minimum, and a uniform focus field is essential. This entails the use of pole pieces of good mechanical design operated at flux densities below saturation.

621.397.62 : [621.396.662+621.396.645 **2033**

Use of New Low-Noise Twin Triode in Television Tuners.—R. M. Cohen. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 3–25.) A full description of the use of the 6BQ7 in 'driven grounded-grid' circuits, whose merits are discussed. Data are given on noise figure, image rejection, gain, and standing-wave ratio, for frequencies in the v.h.f. television bands. The use of the 6BQ7 in an i.f. preamplifier for u.h.f. television receivers is also discussed.

621.397.62 : 621.397.84 **2034**

Locked-in Oscillator for TV Sound.—M. S. Corrington. (*Electronics*, March 1951, Vol. 24, No. 3, pp. 120–125.) The characteristics of f.m. receivers can be improved by decreasing the frequency deviation before discrimination. The instantaneous frequency of the incoming signal is divided by a factor of five by means of a locked-in oscillator. The input to this oscillator must be sufficient to ensure that locking takes place. The method is not suitable for the reception of weak stations with low-sensitivity receivers. A receiver incorporating this circuit is found to give better performance than a conventional receiver when an interfering signal in the same channel is present.

621.397.621.2 : 535.376 **2035**

Saturation of Fluorescence in Television Tubes.—Bril & Kröger. (See 1923.)

621.397.645.37 **2036**

Linearization of the Frequency Response of Wide-Band Amplifiers by Negative Feedback.—Dillenburger. (See 1871.)

621.397.8 **2037**

Investigation of Ultra-High-Frequency Television Transmission and Reception in the Bridgeport, Connecticut, Area.—R. F. Guy. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 98–142.) During the year 1950 systematic comparisons were made between the experimental u.h.f. (530 Mc/s) and commercial v.h.f. (67 Mc/s) transmissions. The receiving stations included numerous homes in which either u.h.f. receivers or converters were installed. The effects of various types of transmission line, aerials and locations are described. The quality ratings of the u.h.f. and v.h.f. pictures were compared and related to field-strength tests made with a mobile equipment. The u.h.f. field strength was more than 20 db below the theoretical level assuming a smooth earth.

621.397.5 **2038**

Television, Vol. 5 (1947-1948) & Vol. 6 (1949-1950). [Book Review]—A. N. Goldsmith, A. F. Van Dyck, R. S. Burnap, E. T. Dickey & G. M. K. Baker (Eds.). Publishers: RCA Review, Princeton, N.J., 1950, 458 & 402 pp., $2.50 per volume. (*J. Soc. Mot. Pict. Televis. Engrs*, Jan. 1951, Vol. 56, No. 1, p. 132.) Reprints of articles by R.C.A. authors.

621.397.62 : 621.396.62.004.67 **2039**

Wireless Servicing Manual. [Book Review]—Cocking. (See 2008.)

VALVES AND THERMIONICS

621.383.012 **2040**

Modification of Current/Voltage Characteristics of a Gas-Filled Photocell with Plane Cathode under the action of a Magnetic Field Perpendicular to the Cathode.—R. Birebent. (*C. R. Acad. Sci., Paris*, 28th March 1951, Vol. 232, No. 13, pp. 1296–1298.) Measurements of the variation of current with voltage were made on an argon-filled photocell, for different values of illumination, with

and without a magnetic field of 250 oersted. The current/voltage characteristic can be represented by an exponential function involving the ionization potential of the filling gas. Application of the magnetic field leads to an increase of current by preventing diffusion of electrons and ions towards the walls.

621.383.27† **2041**

Secondary-Electron Multipliers and their Technical Significance.—K. Nentwig. (*Frequenz*, Dec. 1950, Vol. 4, No. 12, pp. 328–332.) Review of German development, showing details of construction and sensitivity characteristics of new photomultiplier valves. These include an 11-stage valve mounted on a 13-pin base, with sensitivity of the order of 10 A/lumen, though for continuous operation a much smaller current of about 0·5 mA should not be exceeded; for pulsed operation the current may be increased to about 10 mA.

621.383.4 **2042**

Infra-Red Photoconductivity of certain Valence Intermetallic Compounds.—J. G. N. Braithwaite. (*Proc. phys. Soc.*, 1st March 1951, Vol. 64, No. 375B, pp. 274–275.) Short account of investigations of the properties of evaporated layers of Cu_2Te, Ag_2Te, ZnTe, HgTe, Tl_2Te, Sb_2Te_3, Mo_2Te, W_2Te, U_2Te, Zn_3As_2, SnS, and Sb_2Se_3, with a table showing (*a*) the peak wavelength, (*b*) the long-wave threshold, (*c*) the greatest wavelength at which signals were detected.

621.383.49 : 546.683.1.221 **2043**

On the Photoconductivity of Thallous Sulfide Cells.—A. W. Ewald. (*Phys. Rev.*, 15th Feb. 1951, Vol. 81, No. 4, pp. 607–611.) "An investigation of the rise and decay of the photocurrent in Tl_2S cells shows that the simple bimolecular recombination theory on which previous quantitative discussions of the photoeffect in Tl_2S were based is inadequate for the complete explanation of these curves. The dependence of the initial slopes of the response curves on light intensity is that to be expected for a homogeneous photoconductor, but the temperature dependence of these slopes suggests the presence of barriers which influence the conductivity through the mobility. The variations of both the initial slopes and the steady-state photoconductivity with temperature indicate an exponential dependence of the mobility upon temperature with an activation energy of 0·19 eV."

621.385 : 621.318.572 **2044**

Wide-Band A.T.R. and T.R. Gas-Discharge Switches.—R. Musson-Genon, R. Métivier & R. Paliès. (*Rev. tech. Comp. franç. Thomson-Houston*, Nov. 1950, No. 14, pp. 23–34.) The functions and operation of transmit-receive and anti-transmit-receive switches in radar apparatus are described and analysed. Design and performance testing of A.T.R. switches Types RH.0331 and 0332 and T.R. switch Type RH.0531 for use in the X band (8·5–9·6 kMc/s) are considered, and their characteristics are given.

621.385.001.4 : 519.283 **2045**

Statistical Evaluation of Life Expectancy of Vacuum Tubes Designed for Long-Life Operation.—E. M. McElwee. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 137–141.) 1950 I.R.E. National Convention paper noted in 2104 of 1950.

621.385.029.64/.65 **2046**

Transmission-Line Equivalent of Electronic Traveling-Wave Systems.—W. E. Mathews. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 310–316.) The small-signal properties of long electron beams may be analysed in terms of travelling space-charge waves; this suggests an equivalence between such beams and transmission lines moving longitudinally. The analysis of certain electronic devices in terms of coupled distributed-parameter lines in motion is equivalent to a rigorous field-theory analysis. The results for the idealized helix and thin cylindrical electron beam are presented.

621.385.029.64/.65 **2047**

Periodic-Waveguide Traveling-Wave Amplifier for Medium Powers.—G. C. Dewey, P. Parzen & T. J. Marchese. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 153–159.) A theoretical and experimental investigation of singly corrugated coaxial transmission lines is described. The properties of such structures are calculated and the effect of an axial electron beam is taken into account by a field method. Values of gain and bandwidth are derived. The theoretical results agree fairly well with experimental values. An amplifier giving an output of 50 W with a gain of 20 db and bandwidth of 100 Mc/s at a wavelength of 6·5 cm has been produced. The maximum output obtained was 125 W and the highest efficiency 7%.

621.385.029.64/.65 : 621.317.755 **2048**

The Traveling-Wave Cathode-Ray Tube.—H. E. Hollmann. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 194–195.) Comment on 503 of February (Owaki et al.). The prototype of the travelling-wave deflection system is the multiphase system (see 1128 of 1939). The ultradynamic Lissajous figures in the Japanese paper are the same as those obtained by Hollmann many years ago (see 544 of 1940). See also 3536 of 1940 and back references.

621.385.032.2 : 669.198.865 **2049**

Aluminum-Clad Iron for Electron Tubes.—Espe & Steinberg. (See 1949.)

621.385.032.213 **2050**

Boride Cathodes.—J. M. Lafferty. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 299–309.) See 3207 of 1950.

621.385.032.216 **2051**

On Poisoning of Oxide Cathodes by Atmospheric Sulfur.—H. A. Stahl. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, p. 193.) Measurements on oxide cathodes exposed in the open air showed a large decrease of emissivity with increasing sulphur content. Only by sealing the cathodes in a vacuum immediately after manufacture could the formation of BaS (detected by electron diffraction) be prevented.

621.385.032.216 **2052**

The Leaky-Condenser Oxide-Cathode Interface.—A. Eisenstein. (*J. appl. Phys.*, Feb. 1951, Vol. 22, No. 2, pp. 138–148.) "The time dependence of current passed by commercial pentodes has been studied under pulsed conditions of operation. A decay of current is observed in the microsecond time range for those tubes operated for long periods under cut-off conditions. This decay has been interpreted in terms of the resistance and capacitance of the interface layer known to exist from X-ray diffraction studies. Techniques are developed and applied to the measurement of this interface resistance and capacitance as a function of temperature. Following the initial current decay, a partial recovery is observed when long pulses are used. This effect results from Joule heating of the interface layer."

621.385.032.216 **2053**

The Barium-Oxide-on-Tungsten Cathode Interface.—H. P. Rooksby & E. G. Steward. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, pp. 358–359.) The results of

investigations on oxide cathodes lead to conclusions different from those put forward by Hensley & Affleck (263 of January). Compounds similar in structure to the interface compounds have been synthesized by heat treatment of alkaline-earth carbonates and tungstic oxide in the molecular proportions of 3 : 1; they have the formula R_3WO_6, R representing the alkaline-earth metal. Such compounds give X-ray patterns consistent with face-centred cubic symmetry and thus differing from the perovskite structure.

621.385.032.216 **2054**

The Barium-Oxide-on-Tungsten Cathode Interface.—E. B. Hensley & J. H. Affleck. (*J. appl. Phys.*, March 1951, Vol. 22, No. 3, p. 359.) Further investigations confirm the conclusions of Rooksby & Steward (2053 above) that the interface compound is Ba_3WO_6 and not $BaWO_3$ as suggested previously (263 of January).

621.385.032.216 : 546.41 + 546.42 + 546.431]-31 **2055**

Pulsed Emission from the BaO-SrO-CaO System.—L. E. Grey. (*Nature, Lond.*, 31st March 1951, Vol. 167, No. 4248, p. 522.) Dependence of pulsed emission on molecular composition is shown by means of contours on a triangular diagram. A maximum of about 8 A/cm² occurs for the composition BaO:SrO:CaO: :47:43:10.

621.385.15 : 621.385.032.216 **2056**

Secondary-Emitting Surfaces in the Presence of Oxide-Coated Cathodes.—S. Nevin & H. Salinger. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 191—193.) Describes experiments to show that the deleterious effect of oxide cathodes on Ag-Mg secondary-emission surfaces can be overcome by using Ta instead of Ni as the base metal for the oxide coating.

621.385.2 **2057**

The Transit Time, Electron Paths, Cathode Field-Strength and Potential of the Space-Charge Diode for All Values of Initial Velocity, Initial Direction and Current.—A. O. Barut. (*Z. angew. Math. Phys.*, 15th Jan. 1951, Vol. 2, No. 1, pp. 35–42.) The basic equations for the space-charge field in a planar diode lead to a nonlinear equation of the third degree and third order for the x component of the electron velocity-potential. This differential equation and its solution can be used to calculate, without integration of the Poisson equation, the electron transit time, the cathode field, the potential and the electron paths as functions of only two reduced parameters for all values of initial velocity and current under conditions of partial or complete space charge. The generality of the solution is discussed and other possible applications of the method are mentioned.

621.385.2 : 537.525.92 **2058**

The Space-Charge Smoothing Factor.—C. S. Bull. (*Proc. Instn elect. Engrs*, Part III, March 1951, Vol. 98, No. 52, pp. 149–152.) A relation between the anode-current fluctuations and the fluctuation in the total emission of a planar diode is derived; the total-emission fluctations are, in general, much greater than the full shot noise. The space-charge smoothing factor can be expressed in a very simple form and its value can be determined experimentally from measurements of (*a*) the characteristic at two slightly different cathode temperatures and (*b*) the distance from cathode to anode; no measurement of temperature or total emission is necessary.

621.385.2 : 546.289 **2059**

Germanium Diodes.—R. T. Lovelock & J. H. Jupe. (*Wireless World*, Feb. 1951, Vol. 57, No. 2, pp. 57–60.) Description of method of manufacture and short discussion of characteristics and uses.

621.385.2.012 : [546.28 + 546.289 **2060**

Semiconductor Diodes.—(*Electronics*, March 1951, Vol. 24, No. 3, pp. 112–113.) Summary of the characteristics of Si and Ge crystal diodes, with dimensions and terminal data of available U.S. types.

621.385.28/.38 : 537.525.92 **2061**

Measuring the Deionisation Time of Gas-Filled Diodes and Triodes.—K. W. Hess. (*Philips tech. Rev.*, Dec. 1950, Vol. 12, No. 6, pp. 178–184.) A method is described for determining quickly the order of magnitude of deionization time in a triode; an alternating voltage is applied to the anode, while the grid is connected via a resistance to a negative voltage source, the grid voltage giving a direct indication of the number of positive ions present. The cases of a triode in a polyphase rectifying installation and of a diode in a relay circuit are also investigated.

621.385.4 : 621.396.61 **2062**

Two Transmitting Valves for use in Mobile Installations.—E. G. Dorgelo & P. Zijlstra. (*Philips tech. Rev.*, Dec. 1950, Vol. 12, No. 6, pp. 157–165.) Two double tetrodes, Types QQE 06/40 (indirectly heated) and QQC 04/15 (directly heated) are described, in which the screen grids for the two halves form a single structure. When used in an output stage the two halves are preferably connected in push-pull; for frequency multiplication they can be connected in cascade. At frequencies up to 200 Mc/s the QQE 06/40 can generate 90 W with an efficiency of about 75%, and at 300 Mc/s 70 W with 65% efficiency. The corresponding figures for the QQC 04/15 are 22·5 W with over 70% efficiency and 9 W with 34% efficiency. See also *Commun. News*, Dec. 1950, Vol. 11, No. 4, pp. 126–132.

621.385.5 **2063**

Low-Noise Miniature Pentode for Audio Amplifier Service.—D. P. Heacock & R. A. Wissolik. (*Tele-Tech*, Feb. 1951, Vol. 10, No. 2, pp. 31–33, 69.) Design features of the R.C.A. Type 5879 valve are described which result in low values of interelement leakage and of coupling between heater and other parts.

621.385.832 : 681.142 **2064**

The Selective Electrostatic Storage Tube.—J. Rajchman. (*RCA Rev.*, March 1951, Vol. 12, No. 1, pp. 53–97.) The tube is intended for high-speed handling of digital information in the form of on/off signals, and has a storage capacity of 256 signals. Electrons from an extended cathode system are directed to the whole target area, but are prevented from reaching all but one selected window area by means of a complex control-grid system consisting of two orthogonal sets of spaced parallel bars. The characteristics of the tube, the design of the associated circuits, and the arrangement of the bar connections are described.

621.396.615.14 **2065**

The Design of Low-Power Valves for V.H.F. and U.H.F.—R. Suart. (*Radio franç.*, Jan. 1951, No. 1, pp. 1–5.) Discussion of the principles governing the design of triodes and tetrodes for the frequency range 500–1 500 Mc/s. Factors considered include type of structure, electrode connections and disposition, anode voltage and cathode emission. Operating characteristics of seven typical French valves are tabulated. See also 2096 of 1950.

621.396.615.141.2 **2066**

3-cm Magnetrons.—J. Lazzeri. (*Rev. tech. Comp. franç. Thomson-Houston*, Nov. 1950, No. 14, pp. 35–46.) An account is given of the operating principles, design and manufacture of magnetrons, with particular reference to 12-cavity magnetrons suitable for use in

radar on a wavelength of about 3 cm. Characteristics are given of three types, each capable of an output of 40 kW, with an efficiency of about 30%, one with frequency adjustment of 200 Mc/s centred on 9·37 kMc/s.

621.396.615.141.2 : 621.385.833 **2067**

Electron Optical Exploration of Space Charge in a Cut-Off Magnetron.—Reverdin. (See 1988.)

621.396.615.142 **2068**

Amplification by Acceleration and Deceleration of a Single-Velocity Stream.—L. M. Field, Ping King Tien & D. A. Watkins. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, p. 194.) The waves described by Hahn (3521 of 1939) & Ramo (4352 of 1939) not only change in length as the beam velocity changes but also change in amplitude. By a suitable combination of gradual decelerations and sudden accelerations the amplitude may be increased. An amplifier constructed on these lines provided a net gain of 22 db at 3 kMc/s.

621.396.615.142.2 : 621.317.3.001.4 **2069**

3-cm Low-Voltage Reflex Klystron, Type RHK 6332.—J. Chantereau, R. Musson-Genon & R. Métivier. (*Rev. tech. Comp. franç. Thomson-Houston*, Nov. 1950, No. 14, pp. 47–57.) A detailed description of the construction and test methods used. With an anode potential of 300 V, and a maximum anode dissipation of 12 W, the valve gives an output of more than 20 mW over the frequency band 8·5–9·66 kMc/s. See also 683 of March.

621.396.622.63 : 546.28 **2070**

Silicon Crystal Detectors.—J. Mercier. (*Rev. tech. Comp. franç. Thomson-Houston*, Nov. 1950, No. 14, pp. 9–21.) The manufacture, adjustment and performance testing of silicon crystals for use as frequency changers in radar receivers are described in detail. Characteristics are given of a range of crystals for use as frequency changers and as detectors in the 8·5–9·6 kMc/s band and below 4 kMc/s.

621.396.822 **2071**

Transit-time Phenomena in Electron Streams: Part 3—The Electron-Ion Plasma and Beam Fluctuations.—D. K. C. MacDonald. (*Phil. Mag.*, May 1951, Vol. 42, No. 328, pp. 515–522.) A continuation of earlier papers (2408 of 1949 and 1038 of April) on the fluctuations in an electron beam. In this case the equilibrium fluctuations are considered, taking full account of the interaction of the electrons and the field as expressed by Poisson's law. A discussion is also given of 'lattice' models of electron and electron-ion plasma. The noise per unit bandwidth is frequency dependent up to a certain limiting frequency, above which it is constant (pure shot effect). The results are considered in relation to some practical valve problems.

621.385 : [621.396.621 + 621.397.62 **2072**

Receiving Tube Substitution Guide Book. [Book Review]—H. A. Middleton. Publishers: J. F. Rider, New York, 1950, 224 pp., $2.40. (*Electronics*, Feb. 1951, Vol. 24, No. 2, pp. 144, 148.) A list of permissible substitutes for about 750 types of receiving valve is given, with diagrams showing socket changes or adapter construction where necessary. A table of receiving-valve characteristics is included and also a section dealing with television receiver models.

MISCELLANEOUS

061.4(41) **2073**

Festival of Britain.—(*Wireless World*, May 1951, Vol. 57, No. 5, pp. 173–174.) A list of most of the modern radio gear included in the South-Bank and Land-Travel Exhibitions, in the 'Campania' Festival Ship and in the Kelvin Hall, Glasgow.

519.283 : 53 + 621.3].001.5 **2074**

The Control Chart as a Tool for Analyzing Experimental Data.—E. B. Ferrell. (*Proc. Inst. Radio Engrs*, Feb. 1951, Vol. 39, No. 2, pp. 132–137.) 1950 I.R.E. National Convention paper.

53 + 621.3].001.5 **2075**

Research in 1950.—(*Metrop. Vick. Gaz.*, Feb. 1951, Vol. 23, No. 384, pp. 312–319.) A review of the work in various branches of physics and engineering carried out in the Metropolitan Vickers research laboratories, with a list of papers published by members of the staff.

621.3 **2076**

Electrical Transmission of Power and Signals. [Book Review]—E. W. Kimbark. Publishers: J. Wiley & Sons, New York, and Chapman & Hall, London, 461 pp., 48s. (*P.O. elect. Engrs' J.*, Jan. 1951, Vol. 43, Part 4, p. 218.) "A general textbook covering basic theory appropriate to the transmission of power, telephony and ultra-high-frequency signals."

621.3.029.6 **2077**

Ultrahigh-Frequency Engineering. [Book Review]—T. L. Martin, Jr. Publishers: Prentice-Hall, New York, 1950, 456 pp., $6.00. (*J. Franklin Inst.*, Feb. 1951, Vol. 251, No. 2, p. 301.) "It is highly recommended to electronic engineers and physicists who encounter design and operating problems in ultrahigh frequencies."

621.39 **2078**

Radio Technology. [Book Review]—B. F. Weller. Publishers: Chapman & Hall, London, 3rd edn, 420 pp., 30s. (*Electrician*, 6th April 1951, Vol. 146, No. 3799, p. 1131.) Primarily for students of radio engineering. "The chapter on aerials and radiation has been considerably lengthened and a new chapter on u.h.f. technique added."

621.39 Heaviside **2079**

The Heaviside Centenary Volume. [Book Review]—Publishers: Institution of Electrical Engineers, London, 10s. (*Engineering, Lond.*, 26th Jan. 1951, Vol. 171, No. 4435, p. 90.) "The volume contains the papers which were read at the centenary meeting on May 18, 1950."

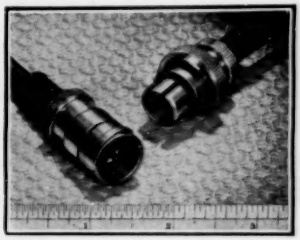
621.396.029.6 **2080**

Short Wave Wireless Communication. [Book Review]—A. W. Ladner & C. R. Stoner. Publishers: Chapman & Hall, London, 5th edn 1950, 717 pp., 50s. (*J. Brit. Instn Radio Engrs*, Feb. 1951, Vol. 11, No. 2, p. vi.) Material no longer topical has been replaced by chapters on sound and vision intelligence, waveguides and wireless telegraph circuits.

621.396.6.001.4.004.67 **2081**

Electronic Equipment Construction. [Book Review]—Publishers: Department of Commerce, Office of Technical Services, Washington, D.C., 300 pp., $7.00. (*Electronic Engng*, March 1951, Vol. 23, No. 277, p. 116.) "This book is a compendium of recent advances in construction techniques and components designed to facilitate servicing and maintenance of electronic equipment. It has been compiled by a team of scientists from the Stanford Research Institute who have visited companies and government establishments both in the United States and England, in order to get the widest coverage on their subject."

The "Belling-Lee" page for Engineers



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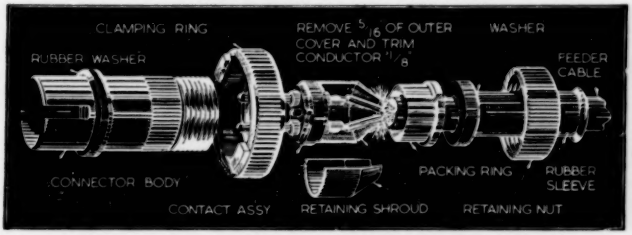

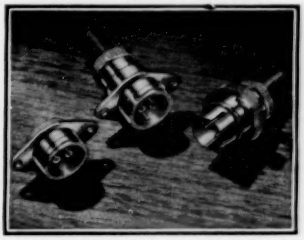

Assembly	Coaxial	2-pole	3-pole
Flex plug	L722/P	L625/P	L715/P
Chassis socket	L722/S	L625/S	L715/S
Through chassis socket	L723	L689	L716
Flex socket	L724	L690	L717

Type	Characteristic Impedance ohms ★	Contact Resistance	Capacitance ★ Conductor/conductor	Capacitance ★ Conductor/screen
Coaxial	75	Less than 5 milliohms each	—	—
2-pole	100		1 pF	2.5 pF
3-pole	—		—	—

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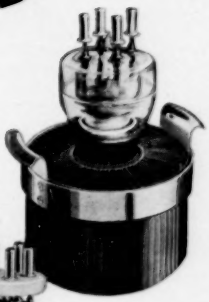

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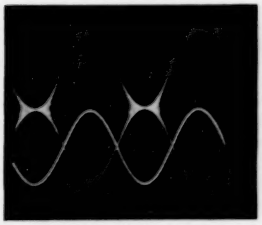
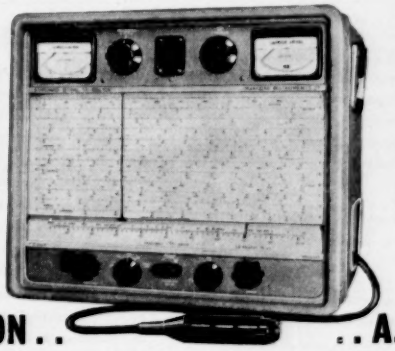
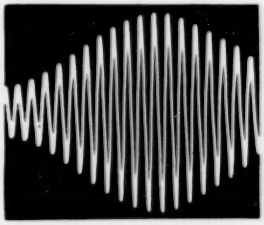


320kc/s modulated 400c/s; audio on second beam.





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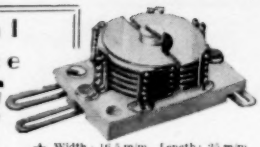

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Apply at once by letter, stating age, full names in block letters, and full particulars of qualifications and experience and mentioning this paper to the Crown Agents for the Colonies, 4 Millbank, London, S.W.1, quoting for post (*a*) M.28660.B, and for post (*b*) M.28666.B on both letter and envelope. The Crown Agents cannot undertake to acknowledge all applications and will communicate only with applicants selected for further consideration.

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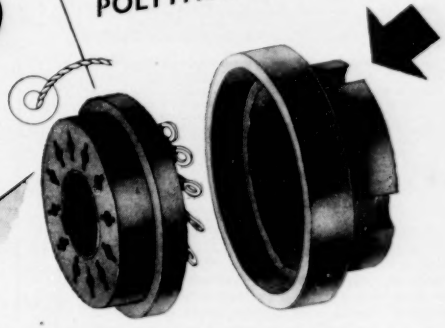

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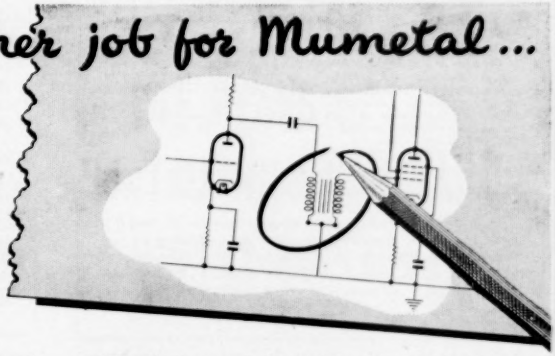

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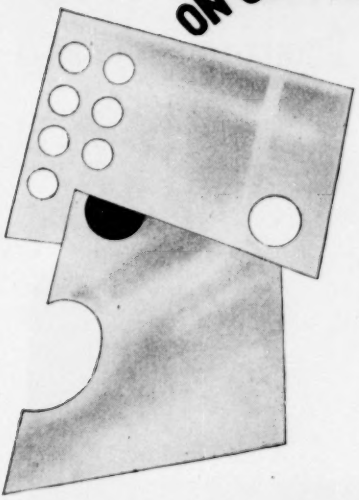

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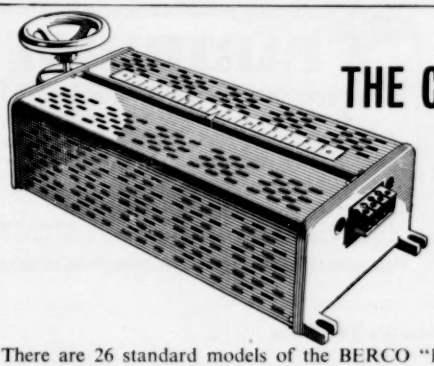

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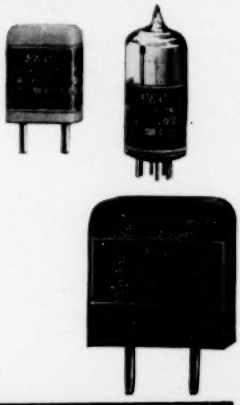

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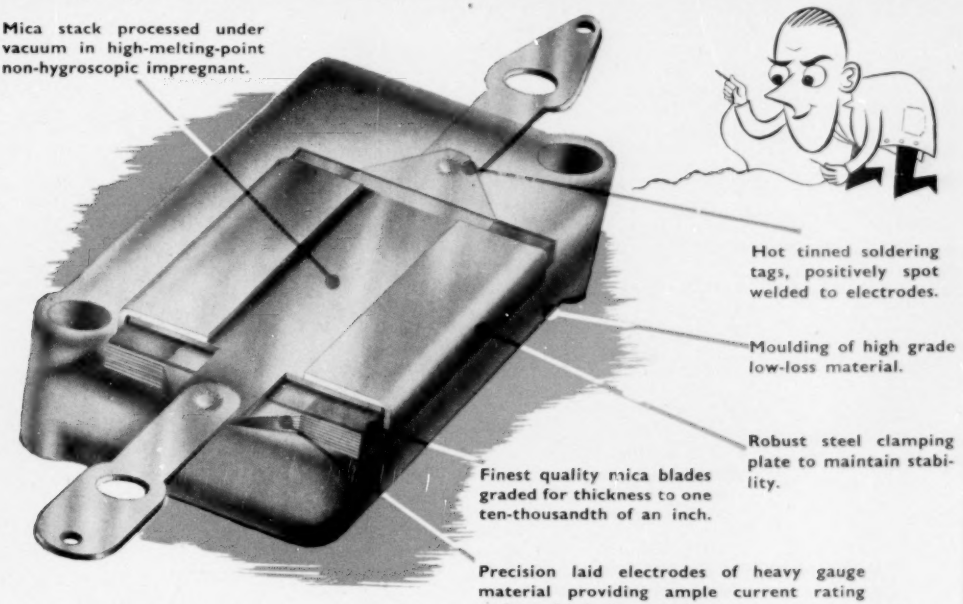
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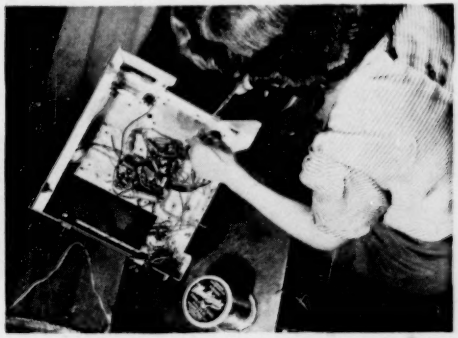

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